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INTEGRATED FUNCTION (CNI) WAVEFORM STUDY

Magnavox Research Laboratories

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FOREWORD

This Final Report was submitted by Magnavox Research Laboratories, 2829 Maricopa Street, Torrence, California, under Contract F30602-69-C-0186, Project 4519, Task 451911, with Rome Air Development Center, Griffiss Air Force Base, New York. Contractor's report number is R-1959. Dean Baerwald, EMCRR, was the RADC Project Engineer for the effort.

To facilitate the dissemination of unclassified information, subsections 7.7.1.1 and 7.7.1.2 of Section 7.7 were removed from Volume II and printed as a separate Volume III. Volume III, Confidential, contains the only classified material in this document.

Distribution of the report is restricted under the provisions of the U.S. Mutual Security Acts of 1949.

This technical report has been reviewed and is approved.

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ABSTRACT

This Final Report, presented in three volumes, describes a comparison of candidate spread spectrum waveforms and the selection of a preferred waveform to perform integrated communication, navigation, and identification (CNI) functions. Satellites are presumed available in appropriate orbits for global communication and navigation. A coordinated frequency/hop/pseudonoise/time hop (FH/PN/TH) waveform is made considering such factors as efficient use of satellite ERP in the remote mode, multiple access of wide dynamic range signals in the direct mode, range and range rate measurement accuracy, initial synchronization, and equipment complexity for full capacity implementation in a nominal 100 MHz bandwidth.

Since CNI system requirements are not presently known, the waveform choice has been made considering a postulated worst case environment based on future air traffic control requirements.

Implementation of the preferred CNI waveform will depend on certain technology developments particularly in the areas of wide dynamic range receivers, phase coherent frequency hopping, high peak power pulse transmitters, and LSI digital devices. However, a demonstration concept can be advanced within the present state-of-the-art to illustrate the preferred waveform with scaled parameters.

Volume I covers the concept formulation studies leading to the preferred waveform and demonstration concept while Volume II summarizes the detailed performance and operational analysis. Volume III presents navigation considerations for the enroute case.

EVALUATION

The purpose of this effort was to investigate all of the various waveforms capable of providing the required information transfer for an integrated communication, navigation, and identification system. The objective was to define the capabilities of the various signal structures and determine the performance trade-offs between them. The approach taken by Magnavox was to define various candidate waveforms and to compare their performance with respect to CNI assumptions and design criteria. In the absence of a definitive CNI system requirement, reasonable postulates were made as deemed necessary for selecting the preferred full capacity waveform and to evaluate performance parameters thereof.

The study results present an excellent analysis of the performance of various spread spectrum waveforms when used in multiple access communications or passive navigation systems, for both point to point and satellite-repeated applications. A frequency hop/pseudonoise/time hop waveform was selected as the best compromise between maximization of theoretical performance and realization of the goal of practical implementation within projected state of the art. An experimental model, using scaled parameters, has been designed to provide experimental verification of the theoretical results. The evaluation of the multi-function concept should be carried out in a laboratory environment before large scale demonstration programs are undertaken.

This study has also identified technology advances needed to implement the CNI concept on an economical basis. These recommendations can be used as a guideline for future programs.

DEAN L. BAERWALD
Project Engineer

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SECTION I

INTRODUCTION

This Technical Report, presented in three volumes, covers the work performed under Tasks A001, A002, and A003 of the Statement of Work, Integrated Function (CNI) Waveform Design, over the period 12 February 1969 to 11 October 1969. The study provides modulation/coding tradeoff studies which lead to identification of a preferred single waveform structure suitable for integration of the communication, navigation, and identification functions into a CNI transceiver/processor satisfying the broad requirements of

- Direct point to point voice/digital communication.
- Remote voice/digital communication through a satellite repeater.
- Passive enroute navigation using satellites in known orbits.
- IFF interrogation and response.
- Landing mode navigation.

Definition of the full capacity waveform with a nominal 100 MHz or wider bandwidth is the major study task A001.

The ultimate CNI system based on the full capacity waveform may be deployed in the 1980 time frame or beyond, and would be implemented with advanced technology to meet operational requirements to be evolved in future system studies by USAF. Within the present state of the art, a demonstration concept is developed under task A002 for the purpose of enabling immediate demonstration of the feasibility of integrating separate functions into a single waveform. The parameters of this demonstration version are appropriately scaled from the full capacity waveform. A functional engineering design of the CNI transceiver/processor implementing the demonstration concept is carried out under task A003.

The study was conducted by Magnavox Research Laboratories (MRL) in Torrance, California, with a portion contributed by the Magnavox Advanced Systems Analysis Office (ASAO) in Silver Spring, Maryland.

SECTION II

CONCLUSIONS AND RECOMMENDATIONS

The Integrated Function (CNI) Waveform Study conducted by Magnavox Research Laboratories has been quite successful in that the concept of a single waveform to perform multiple functions is shown to be feasible and practical. The full-capacity frequency hop/pseudonoise/ time hop (FH/PN/TH) waveform was selected by tradeoff studies as the preferred waveform offering the best compromise between maximizing theoretical performance and realizing a goal of practical implementation within the state of the art. The results of the study are described in a logical conceptual order of defining various candidate waveforms and comparing them with respect to CNI assumptions and design criteria. In the absence of a definitive CNI system requirement, reasonable postulates are made as deemed necessary for selecting a preferred full capacity waveform.

The choice of a frequency hop/pseudonoise/time hop (FH/PN/TH) waveform is made considering the conflicting factors of

- Efficient use of satellite ERP.
- Multiple access of large dynamic range signals.
- Multipath protection with omni-user antenna.
- Synchronization time.
- Range and range rate measurement accuracy.
- Equipment complexity with multi-functions.

Although firm system requirements were not imposed by the Statement of Work for the study, nevertheless it was possible to perform tradeoffs, considering all the above factors, to arrive at a specific preferred full capacity waveform, and to evaluate performance parameters thereof. This choice of waveform will apply to a postulated CNI environment which is believed to be representative of a "worst case". An operational analysis based on information available for typical future air traffic control environment supports this worst case and can be used for more extensive and quantitative evaluations of realistic (as opposed to worst case) multiple access capacities.

Economical implementation of the preferred full-capacity FH/PN/TH waveform at the present state of the art will tend to depend on technology advances in the areas of

- Wide dynamic range RF receiver.
- Coherent frequency hop synthesizer.
- High peak power pulse transmitter.
- LSI implementation (specifically, digital matched filter).

It is recommended that technology developments be pursued in these areas. A transceiver design demonstrating the preferred waveform with the coherent FH/PN precise range tracking loop but with scaled parameters can be specified within the current state of the art for immediate fabrication, or possibly for a restricted breadboard demonstration of critical portions such as the coherent tracking loop.

As possible continuations of the present waveform study effort, a number of areas can be suggested:

1. Analysis on such topics as AGC concepts, synchronization with short code lengths in matched filter, error-correction codes for short messages (IFF).

2. Preliminary design and of specialized sequential decoder for CNI applications (low data rates) using already developed computing algorithm.

3. Further concept formulation studies, such as exotic satellite antenna/waveform combinations, other applications of active ranging, and interfaces with civilian and other military systems.

A full system design is a future step in the orderly evolution of the overall CNI implementation.

This report is presented in three separate volumes. Volume I covers the assumptions, the concept formulation effort leading to selection of a preferred waveform, the technology implications thereof, the demonstration concept, and the conclusions and recommendations of the study. Volume II presents the results of performance analysis and simulation and the operational analysis, which support the waveform choice given in Volume I. Volume III presents navigation considerations for the enroute case.

SECTION III

ASSUMPTIONS AND DESIGN CRITERIA

The major assumptions and design criteria that are used in the waveform study, so as to yield an integrated CNI design that efficiently meets the requirements in the RADC statement of work, are listed as follows:

3.1 ASSUMPTIONS

1. Though the study covers modulation tradeoffs rather than overall system design, the requirements imposed on system interfaces by each waveform need to be considered. Examples would be those concerning the need for satellite processing, if any, steerable antennas, and ILS transponder configuration.

2. An omni-antenna is assumed in the aircraft for the remote satellite mode as well as the direct mode. This represents a system worst case and is to allow for inclusion of small, lower cost aircraft in the system. Large aircraft users, which could accommodate directive antennas in the satellite mode and which, therefore, represent a best case, could obviously be compatible with the worst case waveform design. Carrier frequencies in the lower end of the 1 to 10 GHz band are assumed.

3. A four-satellite, 621 B (NAVSAT) geometry is assumed for passive hyperbolic enroute navigation.

4. No specific COM satellite configuration is presumed here, except that plausible satellite parameters are anticipated (satellite maximum RF power amplifier output of 200 W to 1 KW average; multiple electronically steerable, 2-degree pencil beams in addition to an earth coverage beam; the COM satellite is at synchronous altitude). Different waveform designs will, however, constrain the number of COM satellites that can be used simultaneously in a given field of view from an aircraft terminal and in a frequency band (this pertains to coordinated waveforms versus uncoordinated waveforms). Nevertheless, if one presumes adequate isolation between satellites, on the basis of either antenna beam spatial discrimination or frequency discrimination then a waveform solution that is designed for all COM traffic through a single satellite, is also applicable to the case where multiple satellites are

simultaneously used for COM. The latter would include the case where each of the four satellites in the 621 B configuration are simultaneously used for COM and NAV.

5. In the ILS mode the beam rider and active ranging concepts are presumed. However, should the system design ultimately use hyperbolic ILS navigation instead, the waveform design for the enroute phase would still apply.

6. In the IFF mode, no one IFF system concept will be presumed. However, the capability of the candidate waveforms to be used in the alternate concepts will be considered (alternate IFF system concepts include addressing and reply on the basis of known aircraft position, addressing and reply on the basis of exclusive slot assignments that can be made in a highly disciplined system, and addressing on the basis of known angular orientation and discriminating the replies on the basis of time-of-arrival).

7. Though no scenarios have been furnished with the Statement of Work (that would give COM link needlines, near-far power distribution, multipath geometries, and jam threats), nevertheless, a common bounding set (including worst case and average case) of assumptions can and are used to evaluate all modulation techniques. For example, the worst case near-far situation for multiple access in the direct mode would have all other access signals received at greater signal strength, whereas the average case situation would have half the accesses at greater strength and half at less strength. This would be for a waveform concept involving no power control. The multipath reflections for the different accesses are also weighted by the above near-far power assumptions as well as by the expected reflection characteristics for a given earth situation. In the case of jamming, both the simple broadband noise type (for which the modulation processing gain is the key factor) and optimum intelligent jam type are considered. In the latter regard, since difference optimum jam strategies apply to different modulations, then different jam threats are considered.

8. Though common band operation of all aspects of CNI is desirable, including the direct and remote modes, the obvious use of separate frequency

bands as appropriate, is assumed . still part of this study. A bandwidth in the order of 100 MHz is considered for each band so as to achieve at least a theoretical 46 db processing gain for a 2400 bps channel.

9. Though the availability of precise global timing and of atomic clocks is not presumed, it is assumed that operational procedures to allow transfer and update of timing are allowed. The airborne terminal then can be postulated (for purposes of the study) to have a timing accuracy that corresponds to the precision of measuring range actively, or the precision of measuring NAV using the NAVSAT's, even though the aircraft may have only a 10^{-7} clock. Similarly, insofar as interface with adjacent terminal areas is concerned, any necessary coordination (e.g., in slot assignments for coordinated waveforms) is not considered to be a limiting factor.

10. The aid of an inertial platform (for NAV or sync tracking) is presumed not necessarily available. However, utilization of its availability is considered.

3.2 DESIGN CRITERIA

The overall design goals for the CNI waveform are to reduce the plethora of black boxes (and thus complexity) by means of commonality of waveform design, to extend performance in critical areas over the present state-of-the-art, and to maximize overall performance effectiveness. In these regards, key design criteria that are considered in the design of candidate systems for CNI include the following:

1. Though a common waveform is the desired goal and though the basic RFP requirements can be met with a given design, typically the relative performance efficiency for each of the COM, NAV, and IFF function modes of a design is different. Therefore, the degree to which one should be optimum, say in the COM mode versus being optimum in the NAV or IFF mode, must be considered. Depending on the relative weighting, such tradeoff deviations from commonality may be considered.

2. Similarly, since the environment for the direct mode is significantly different from that for the remote mode, the optimality of the common waveform design for the two modes must be traded off and appropriately weighted.

3. Dynamic range (which means the capability to receive, in the nearby vicinity of independent transmitters, a desired signal from a distant transmitter) should be maximized. This factor (sometimes described as the near/far problem), involves both the relative distances and power differential involved (greater than 100 db is a plausible worst-worst case, 80 db is a plausible worst case, and 30-40 db may be a plausible average type case), and also the number of adjacent transmitters.

4. Multipath interference (considered both in terms of forward propagation and also in terms of backscatter propagation) represents a significant interference situation that should be discriminated against. Forward scatter multipath could limit detection reliability and accuracy, whereas backscatter multipath could limit dynamic range.

5. Spectrum splatter and time splatter, due to equipment effects, and pulse shapes, should of course also be minimized. The effects are particularly significant in limiting dynamic range.

6. System power and bandwidth efficiency should be maximized. This allows both a maximization of the number of simultaneous multiple access signals allowed and the amount of A/J obtained, for the cases of COM and IFF. For the case of NAV, both the accuracy and fix rate are maximized with or without jamming. There is a plausible tradeoff between the two factors of antijam and multiple access. Selecting a design that maximizes this tradeoff curve for all values of the parameters is, of course, desirable (e.g., coordinated waveform systems). On the other hand other system designs (uncoordinated ones) may not allow optimization at all values. For example, a modulation/coding combination that maximizes multiple access performance in the direct mode (with a near-far problem) in the absence of jamming may not be the optimum combination against jamming in the absence of multiple access. Therefore, in this case a weighting of the significance of multiple access efficiency versus antijam efficiency must be considered.

7. The capability to obtain accurate NAV measurements under the conditions of high range acceleration (e.g., 8 g's for high-performance aircraft), and also high range rate or Doppler, should desirably be obtained. Otherwise, there would be a tradeoff of having to depend on external aids, such as an inertial platform.

8. The connectivity of the system should be enhanced by appropriate waveform design. Connectivity denotes the ability of any of a group of desired terminals to be connected (at least in a monitoring and interrogation sense) to all other desired terminals at the same time. A tradeoff is, of course, involved here in the number of parallel equipments one should provide with any resultant limitations on dynamic range versus the impact on other criteria due to the constraints imposed on the waveform.

9. Total system reaction time to a change must be considered and minimized as well as the time to acquire sync in the receiver. Actually reaction time is the sum of sync time and time for supervisory signaling and parameter setting when the system must be reorganized. For uncoordinated waveform systems the reaction time is essentially just the sync time, whereas for coordinated types of waveform systems the reaction time includes the additional operations plus the relative position of one's slot in a frame (for time-hopped waveforms). The relative significance of given values of total reaction time must be considered.

10. The design values for sync acquisition time should include both the corresponding value of probability of success expected as well as the length of time involved (20 msecs in statement of work). Failure to achieve sync in that time means that repeated tries must be made, and, therefore, more time taken. In the case of time hopped waveform designs, where a preamble is allowed for sync, this can further mean loss of a block of data in addition to the fact that repeated sync acquisition attempts must be made at intervals spaced by a whole frame length.

11. The full utilization of the information in each of the three channel outputs (COM, NAV, IFF) should be used to aid the operation of the other channels (the airborne computer is also employed in the processing) so as to maximize total performance for each function. Different waveforms will have different capabilities in this regard that will impact, therefore, on total performance.

12. The particular satellite system, antenna system, and ground station system that is assumedhas a significant impact on waveform performance, and, therefore, choice. One cannot really design the optimum CNI waveform without considering possibilities in these regards, such as satellite processing

schemes and multiple narrow beam antennas. In comparing alternate waveforms, the differences in interface requirements must be factored in, though of course the impact on the airborne transceiver should be of dominant importance.

13. The waveform design should have flexibility. That is, though the system would meet the current requirements of the statement of work, it should not be a rigid design that is frozen to these requirements, with a significant redesign needed to allow a requirements change. For example, plausible changes one could conceive are a change in bit rate (and possibly bit error rate) to allow compatibility with other communication systems (e.g., vocoder versus delta mod voice), increase in multiple access requirements as part of normal and projected system growth, change in direct mode range (e.g., 600 miles versus 300 miles), change in frequency spectrum assignment or occupancy, change in aircraft antenna and satellite system types one interfaces with, and change in ranging or navigation technique (this is meant to be the philosophy or concept of ranging; e.g., transponder methods versus hyperbolic ones - and it is meant in the sense of changes from the initial assumptions of the study).

14. Maximum compatibility with a civilian environment is desired. This is taken to mean that more economical, but lower-performance, civilian derivatives of the military design can be synthesized that are effective, and that the military design must be able to coexist with current civilian systems (e.g., collision avoidance and landing systems). As far as being able to evolve gradually from current systems to the ultimate military system, this is not considered to be an initial constraint in concept formulation here. Rather, it is a factor that one considers after an optimum design has been determined. It is obviously a desired goal that a gradual evolution be possible. However, the weighting of the overall significance of compatibility with a civilian environment is not really considered here.

15. Cost and complexity, and time frame of implementation are also to be considered as significant criteria in waveform system synthesis. Cost and complexity factors are considered in a relative and qualitative sense. However, the ability of a given waveform design to allow timesharing of equipments has a more specific measure. The weighting of performance versus the factors of cost and complexity is significant to waveform selection. Time frame of

implementation of an optimum system is considered to be in the next decade or so, but a technological base to insure implementability of all the techniques in the early 1970's is desirable, at least for purposes of demonstration feasibility.

16. Digital processing techniques will be emphasized for implementation since they have potential to benefit from LSI/MSI computer technology. Digital filters, Fast Fourier Transform, digital matched filter, digital phase lock loops represent examples which could be implemented either as special purpose digital circuits or by computer software.

SECTION IV

CONCEPT FORMULATION

In this section the overall waveform design concepts, that meet the requirements of the Statement of Work and are designed in accordance with the assumptions and criteria stated in Section III, are considered. This discussion of waveform concepts includes identification of the basic alternate waveform types considered and the rationale for their consideration, a comparison of their relative performance, the selection of a plausible one of these for more detailed study, and a detailed description of this waveform in terms of its overall performance, operation, and interfaces. The scope of the concept study here is solely within the A001 Task of the Statement of Work. Demonstration concept aspects (within the scope of the A002 and A003 Tasks of the Statement of Work) not considered in this concept formulation discussion are treated in Section VI. Also, it is noted that the key techniques involved in the waveform design are only identified and discussed briefly as appropriate to the overall concept formulation in this section, but that the detailed analysis and synthesis of techniques and also the detailed performance analyses are covered in Volume II.

A basic waveform discussion follows.

4.1 GENERIC WAVEFORMS

The basic or generic waveform types are categorized here in terms of coarse and fine structure. The coarse waveform structure is determined by the extent and nature of coordination that is designed among the different access signals. The fine waveform structure is determined by the access signal characteristics. Though many waveform varieties are still possible with this type of categorization, one can limit the list to five broad types that representatively emphasize different criteria in Section III and that all meet the requirements of the Statement of Work (but with different degrees of overall performance effectiveness since different criteria are emphasized). These five types are identified by their salient characteristic, as discussed subsequently. Many additional waveforms are, of course, possible but these are all considered as variations (such as in modulation/coding combinations, or in the specific blending of hybrid characteristics) in the appropriate category types. By examining the illustrative waveform types initially, one can determine both the best ones as well as the specific variations that optimize performance.

The five waveform system types comprise:

1. A coordinated, frequency hopping waveform - the COM, NAV, and IFF functions each use similar frequency hop waveforms to the maximum extent possible but they can be time division - or frequency division multiplexed with each other.

2. A coordinated time hopping waveform - the COM, NAV, and IFF functions each use this type of waveform and are also time division or frequency division multiplexed with each other.

3. An uncoordinated frequency hop waveform - the COM, NAV, and IFF functions are also uncoordinated with each other and use similar waveforms.

4. An uncoordinated time-hopped waveform - the COM, NAV, and IFF functions are also uncoordinated and use similar waveforms.

5. A hybrid involving a coordinated waveform; e.g., a time-hopped one, in just the satellite mode but a corresponding uncoordinated waveform in the direct mode.

A pure pseudonoise, spread spectrum, multiple access waveform is not included as a basic waveform type because its fundamental limitations to handle dynamic range in the direct mode rule it out of consideration here. Now in each waveform type indicated, a hybriding with other waveforms is included where desired, but the dominant characteristic of the waveform is the one used to describe it. For example, the FH/PN waveform that involves frequency hopping of low-chip rate PN waveforms is described as predominately an FH waveform type; a TH/PN or TH/FH waveform that involves only a few PN chips or a few frequency channels, respectively, is described as predominately a TH waveform type. Also, in the case of hybriding methods of coordination, if the bulk of the traffic capacity (i.e., most of the 2400 bps COM accesses) is coordinated, then though the NAVSAT or some signaling or IFF functions may be uncoordinated, the waveform is described as predominantly a coordinated one.

Also, it is noted that all five waveform types here are considered to be random access, discrete address waveforms (the coordinated waveform types would need a

discrete address signaling function for this purpose whereas the uncoordinated waveforms have this as an inherent by-product).

4.1.1 COORDINATED FH

The coordinated frequency hop waveform is one where each user is coordinated, by means of time sync and possibly NAV platform and power control aids, so that the transmissions will overlap in frequency only for a small percentage of a bit interval, if at all, so that coding can correct for this overlap and so that each user transmission is detected essentially free of mutual interference. The frequency hop sequences used by each terminal together occupy the entire band. Because of the limited number of accesses and the desire to provide added protection against multipath and jamming, each user waveform is also modulated by narrow band PN. One then actually has a coordinated FH/PN waveform where the band occupancy is flat for all users at the same power level. Such a waveform would be used for the COM and NAV and disciplined IFF transmissions. However, for discrete address signaling that precedes getting into the coordinated state for COM and for random access IFF a swept frequency or spectral analysis type of implementation is required if the transmissions here are to be realized as essentially non-overlapping. Alternatively, uncoordinated FH/PN waveforms would be used for the IFF and discrete address signaling functions. Error correction coding would be depended on to reduce the effect of mutual interference. Because of the smaller number of signaling accesses (and random IFF) the use of such uncoordinated waveforms would not degrade system efficiency as much as would be the case if such waveforms were used for the COM function, which have a large number of accesses.

Now for the coordinated FH/PN waveform system here (where it is seen that this description pertains primarily to what waveform is used for the high capacity or COM traffic, regardless of whether coordinated or uncoordinated waveforms are used for the IFF and signaling functions), the COM, NAV, and IFF functions are further coordinated by sending them on a time division multiplex basis. That is, in a given frame interval a separate time slot is assigned for NAV transmission and fixing (actually this slot comprises four sub-slots to allow for fixing from four separated NAV signal sources), another time slot is assigned for random IFF and discrete address signaling for COM (the time slot actually consists of the three sub-slots to allow handling of three simultaneous queries), and the rest of the time is assigned for COM

in the coordinated state. The NAV and IFF/signaling slots occupy a small percentage of the frame because of the low information content for these functions.

The reasons for such TDMA control of the COM, NAV, and IFF functions are fourfold: to minimize mutual interference between functions, and thus, to maximize efficiency; to allow timesharing of receiver (and possibly satellite antenna) equipments; to avoid a sampling loss in link S/N if one were to time share equipments without coordination; and to provide for high track loop bandwidths for the low-bit NAV and IFF functions so that large range accelerations due to aircraft maneuvers will not result in loss of lock. To achieve rapid sync for each function, and also to allow for effective discrete addressing, a digital matched filter for the PN component is included as part of the FH/PN waveform concept here.

Consider further the description of the coordinated FH/PN waveform approach by discussing separately each of the COM, NAV, and IFF functions, the method of sync, and the multiplexing of the functions.

4.1.1.1 Communication

For the COM function, a 200 KHz PN-coded waveform is used on any of 500 frequency channels that together occupy a 100 MHz total bandwidth. This allows 500 simultaneous COM accesses on separate frequency channels; i.e., with no co-use of a frequency channel. The PN spectrum and frequency channel selectivity is so designed that negligible spectrum splatter and cross-talk occurs, so that in essence, orthogonality of waveforms exists even in the presence of a "near-far" power differential at least as high as 80 db (a tunable RF preselector is required). Frequency hopping of channels by each user also is provided to provide spread spectrum protection against jamming and diversity against multipath. The hopping is coordinated amongst the 500 accesses (the game of "musical chairs") so that orthogonality of channel usage at any instant is maintained (except for propagation delay overlap) and all 500 channels are occupied at any instant. To a jammer, the system hopping still appears to be essentially random. With such an orthogonal waveform design, a high multiple access/AJ efficiency results (the AJ protection is actually independent of the fraction in use of the 500 total maximum access capacity).

Because of propagation delay differences (maximum of 1.8 milliseconds for 300 miles) amongst COM terminals, and of course the fact that omni antennas make one

vulnerable to mutual interference unless one designs against it, the maximum coordinated hopping rate is limited. If, for example, it is selected as 150 hps (for a hop dwell time of 6.67 milliseconds) then there could be a mutually interfering overlap of up to 25% (1.8 milliseconds out of 6.67 milliseconds) from a distant terminal that had just been using the same frequency. Increased overlap of course results from a higher hop rate, and less overlap from a lower rate. By providing an additional 25% redundancy (i.e., over and above that for the redundancy in the normal coding design for direct in-channel interference) in the number of digits per bit or in guard time, the 150 hps is allowable without degrading the orthogonality of waveform design for the coded information bits.

In discussions of the coordinating users problem, reference is typically made to "nets". It is assumed that a net comprises a particular frequency hopping pattern, and only one user in a given net is transmitting. A given terminal may be in several nets simultaneously, such as a ground station controlling several groups of aircraft. Either coordinated (orthogonal) or non-coordinated operation of the nets may be presumed. Each net may be viewed as a single COM access.

For an aircraft initially to get into a designated frequency hop net of the orthogonally organized and coordinated COM accesses, where the number of slots is less than the number of addresses, the design here would use random access discrete addressing. This would be done in a separate allocated time slot which is the same one as for the IFF function which will be described later. (This is a 20-60 millisecond time slot every 1 second). The same FH waveform is used as for the organized message transmission portion of COM, but since it is uncoordinated, the hop rate is made 2400 hps or even higher as desired. The discrete addressing, for the case of an aircraft, will be directed to a central ground station in the airport terminal area (could be the same master control station used for the hyperbolic NAV transmitters), which by monitoring and use of stored information, can rapidly determine the availability of and then make initial assignment of an orthogonal frequency hop sequence for aircraft use.

For a simple illustrative modulation/coding combination, the COM channel would use DPSK modulation together with double coding (triple repetition with 2 out of 3 needed for detection, and 23, 12 Golay coding). The 6 to 1 redundancy in this coding

together with the additional 25% indicated previously for frequency hop switching, gives a total of 8 to 1. For a 2400 bps VOCODER rate the digit rate is 19,200/second; hence, there is 10 db of PN processing for each digit at the 200 kbs PN chip rate. The same modulation/coding combination and digit rate is used for the random access signaling. The processing gain (100 MHz divided by the 2400 bps rate) is 46 db for the COM message signal, whereas it theoretically can be greater than this amount by the reciprocal of the time slot duty cycle (by 14 db) for random access signaling, assuming that a peak-to-average power trade is realized and that the location of the time slot in each frame is pseudorandomly permuted.

The modulation/coding combination indicated gives an E_b/N_0 of 11.5 db for a 10^{-5} bit error rate in additive Gaussian noise. More efficient techniques that one would use include sequential decoding and even combination of this coding with a more efficient modulation such as coherent PSK. Rate-1/2 sequential decoding when combined with DPSK would reduce the required E_b/N_0 to 7 db as described in Section 7.3.1 of Volume II. When combined with coherent PSK instead and with quantized demodulation, the theoretical E_b/N_0 is down to 3 db, but the practically implementable value would be about 5 db. However, the factor should be weighed that the cruder DPSK modulation/coding combination is relatively simple and does not require compensation inputs from the NAV channel for efficient coherent detection while the aircraft is maneuvering. In contrast, coherent PSK track loops for low bit rates need compensation for they would otherwise lose lock during significant aircraft accelerations.

Because of the long hop dwell time (400 microseconds or 80 miles of propagation delay for a 2400 bps rate) one would not expect multipath reflections to overlap other transmissions and, thus, cause mutual interference. Instead, multipath will overlap one's own transmission. For multipath delay greater than a PN chip interval (5 microseconds or 1 mile) it will cause intersymbol interference, but is attenuated by the 10 db PN processing gain, so that error correction coding should be able to handle the residual effects. For multipath delay of less than the PN chip interval, the effect will be that of "slow and flat" fading. However, error correction coding will still correct for fading errors - in fact, the multipath energy tends to be used constructively. The effectiveness of the error coding here, of course, depends on random independent errors on a digit by digit basis. However, a fast frequency hopping over channels spaced by more than the reciprocal of the differential multipath delay, will decorrelate

the fading errors at the frequency hop rate so that the amount of bit interleaving needed is small, to assure random independent errors as required for error correction by sequential decoding.

Figures 4-1 and 4-2 indicate the effect of multipath which is assumed to have a Rayleigh distribution (i.e., diffuse). Figure 4-1 indicates the effect of multipath fading on the FH/PN waveform approach for COM versus the average relative power level of the multipath reflection, when the fading rate is slow compared with the digit rate. Figure 4-2 shows the corresponding effect when long multipath causes intersymbol interference. Also, superimposed on the curve is the result when just a fast hopping (at the digit rate) is used to hop away from multipath so that PN encoding is not needed. The curve in Figure 4-2 is plotted versus the average relative power level of multipath as seen after PN processing in the receiver (for equal strength multipath into the receiver, it should be considered as 10 db down for use of the curves in Figure 4-2).

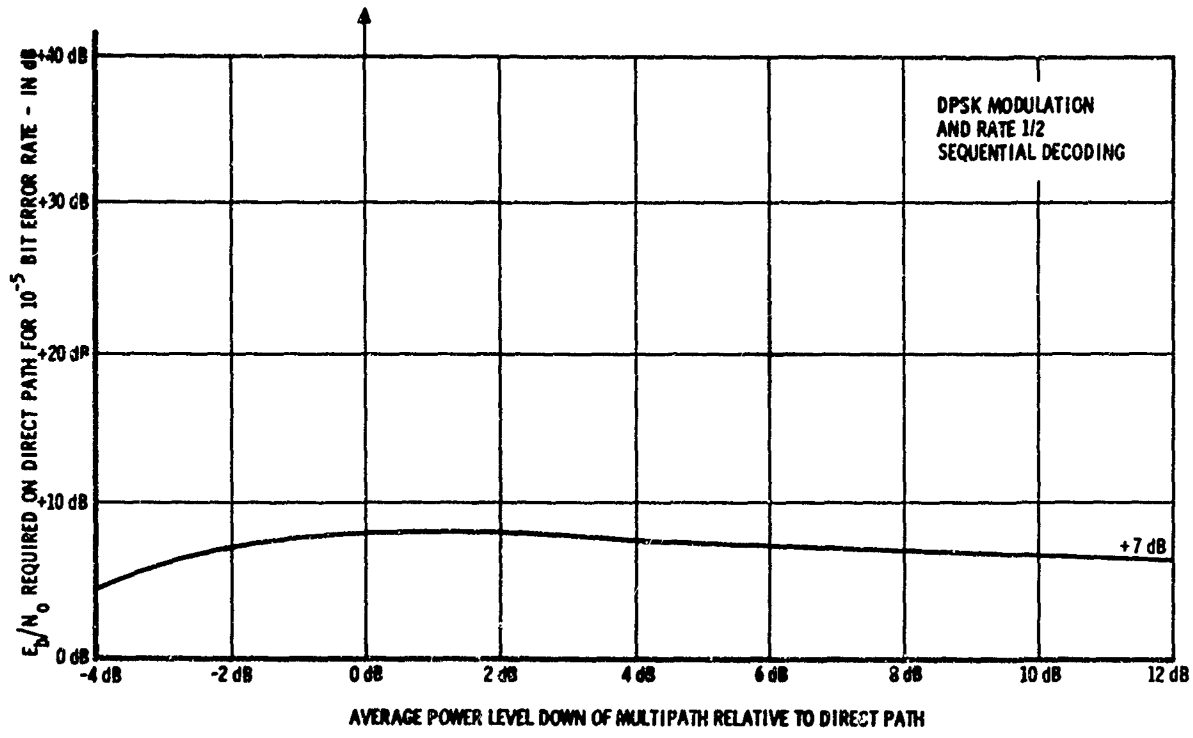

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Figure 4-1. Effect of Multipath Fading (With Pulse Overlap) on COM Performance.

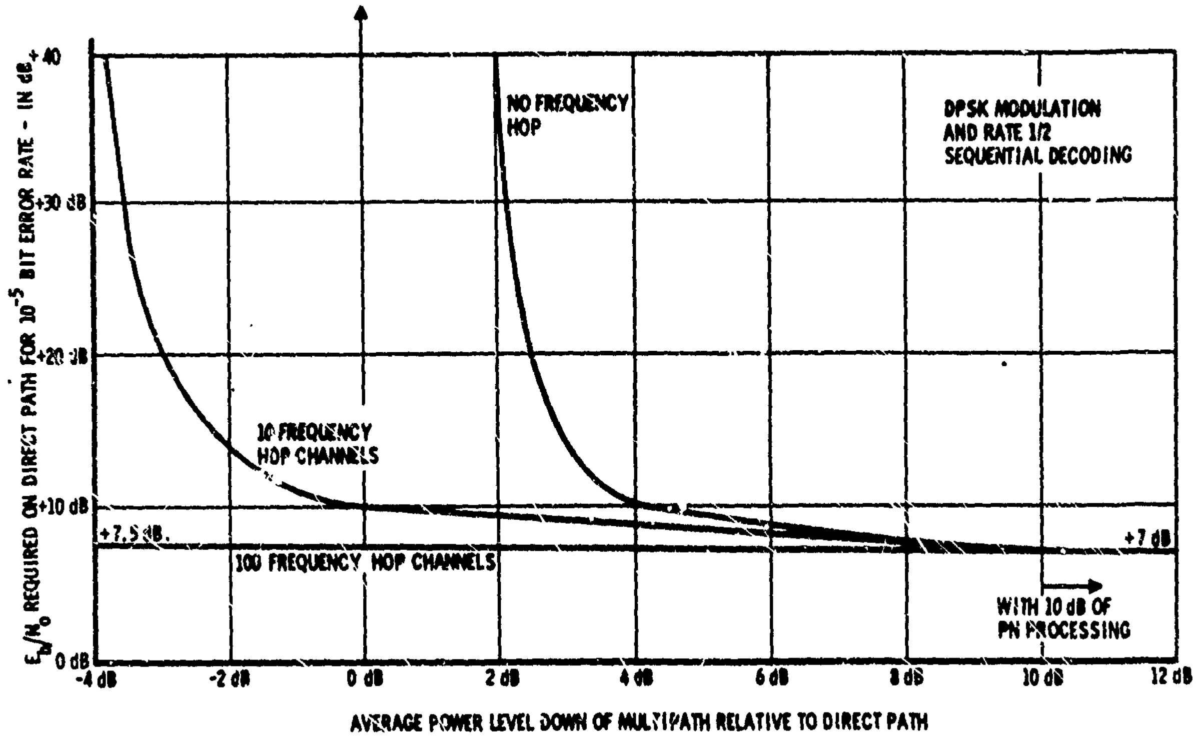

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Figure 4-2. Effect of Multipath Intersymbol Interference on COM Performance.

Both Figures 4-1 and 4-2 indicate the results with DPSK modulation when combined with rate-1/2 sequential decoding, as discussed in Section 7.3 of Volume II. It is seen from Figures 4-1 and 4-2 that multipath of average power up to the direct signal has a negligible impact on system COM performance for the FH/PN waveform approach (at most 2 db).

The above discussion on multipath effects has been concerned only with the forward propagation multipath, which causes fading and/or intersymbol interference effects. Backscatter multipath, though down in level but within the dynamic range limits of interest, is not discussed because the dwell time, at a given frequency, of each aircraft transmitter is greater than the 200 microsecond maximum spread of backscatter multipath at 100,000 foot altitude. Therefore, this is not a problem area for this type of waveform design.

For operation with a hard-limited satellite repeater, the FH/PN waveform concept requires power control of all uplink COM transmissions to the satellite. For

just CNI users, whether they be small aircraft or large aircraft or ground terminals, this should be no real problem (except for deep multipath fades) provided each channel per CNI user takes the same 2400 bps information rate. A control scheme is, of course, still required for each transmitter to compensate for power fluctuations due to propagation range and antenna pattern variations. If desired, the CNI satellite system master control station can process all the status information made available to it and direct the power settings for each transmitter. One has here the tradeoff of central control versus adaptive self-control. The maintenance of control need be accurate only to a few db. On the other hand, if the COMSAT in use is to be more of a general purpose satellite, rather than just a dedicated CNI satellite, then the possible mix of dissimilar terminal users of the satellite creates additional power control problems. These are solved for the FH/PN waveform system here if channelized processing is provided in the satellite for the CNI system users, such as designed in current TACSATCOM concepts. With regard to intermod generated by multiple FH/PN accesses through a hard-limited satellite repeater, the results should be more favorable than if pure or discrete frequency hop signal waveforms were used. The reason is that the FH/PN accesses, which are essentially of equal power, and which uniformly fill the rest of the band, should appear as Gaussian noise processes.

Because the satellite links are power starved for the case of omni aircraft antennas, the system has increased vulnerability to up-link jamming, where the jammer would be stealing satellite power, thus, increasing his effectiveness. For this reason, it may be desirable to provide satellite processing in the form of frequency dehopping for the FH/PN signals. A tradeoff exists between the need to provide such modulation processing on-board versus the AJ discrimination possible with very narrow satellite antenna beams on the up-link and on-board antenna processing. In any case, as far as the modulation system is concerned, the use of FH/PN is readily compatible with on-board modulation processing, as described, if one sought to use it.

4.1.1.2 Signaling and Random IFF

For random access IFF (as distinct from a disciplined IFF system design that would use the same waveform design as for COM), the aircraft's address is given by his instantaneous NAV position together with code control from a common time standard. The interrogator presumably knows the NAV position (as determined by other

sensors) of the aircraft he is querying, and, of course, has the code-of-the-day and the same time standard. Random access, discrete addressing with nonorthogonal waveforms of the queried aircraft is presumed. For example, if one wanted to resolve 0.1 mile intervals in space so that 10^9 NAV addresses are involved in an airport terminal volume, then since only 500-5000 orthogonal addresses are available by frequency channeling (for conditions of a near-far power differential, only frequency channeling or time channeling can be considered as giving real orthogonality) non-orthogonal situations could result depending on the aircraft deployment. If it is not required to resolve aircraft with such precision, the design is, of course, alleviated somewhat. However, a large number of required addresses are readily obtained in this system by the fact that the FH/PN combination gives up to 2^{10} different combinations (with 10 db of PN processing) for each of the 6 redundant pulses per bit, all of which could be sent on any of 500 different frequency channels (a 2400 bps rate is assumed in the IFF mode - if instead one hopped at the digit rate, then the number of address combinations is increased since there now are 500 different channels possible and independently selected for each of the 6 pulse positions). This number is clearly much greater than 10^9 so that no problem in numbers of addresses exists. The interrogation message contains sufficient bits to indicate the instructions involved, identification of the interrogator, and even possibly the location of the interrogator (though typically the latter information would not be included). The interrogated aircraft replies, using its rf address that is based on NAV position, but including in the reply sufficient information bits to indicate identification and position coordinates. The latter would require 27 bits (for 0.1 mile precise coordinates, and therefore 10^9 coordinates in 300 mile x 300 mile x 15 mile space), whereas the former would require less than 20 bits. Allowing a few bits for special messages gives a total of 50 bits needed for the IFF message considering the factors just discussed.

Authentication of the IFF query as well as of the IFF reply is accomplished using the same pseudorandom key generator as provided for the COM function (one keys at the bit rate and the same bit rate is used for IFF and COM) for military users since the functions are time multiplexed and equipment can be shared. Authentication of random access signaling that sets up the COM function, would also be accomplished for IFF using the same key generator. Synchronization of the key generator (which runs continuously after being set at take off) is maintained by the common time standard on board each aircraft (the time standard must maintain key-stream bit integrity to

less than 0.5 bit or 200 microseconds for 2400 bps) so that for each IFF message additional bits need not be sent for resync. If a key generator had not been available, as would be the case for civilian users, then inclusion of extra bits in the IFF message can still be avoided by an alternate authentication technique. This would be to measure the time-of-arrival (using NAV equipment) of the IFF reply and correlate it with the known (from other sensors) position of the queried aircraft.

Since the same bit and digit rates are to be used for IFF as for COM, and also the same modulation/coding is used, a 50 bit IFF message requires a 20 millisecond time slot. To allow for reception of multiple IFF queries and also allow for transmission of an IFF response in any given frame, 3 of these 20 millisecond time slots are allocated.

Actually, the IFF function slotting as just described, would be used for either random type IFF or random access signaling purposes. The significance of designing for multiple time slots, and even of designing to time slot the entire IFF function with respect to the COM and NAV functions, is that equipment such as receivers and transmitters can be time-shared. (In contrast, if equipment were designed for FDMA stacking of the COM, signaling, and IFF functions or multiple IFF channels, a proliferation of equipments would result.)

The multiple access capacity for the IFF and signaling function has been evaluated here for the case where thermal noise is neglected and for the worst case where all other interferers have a stronger signal because of the "near-far" power differential in this worst case deployment. The results obtained for non-coordinated access in one sub-time slot must be multiplied by three, since there are three such sub-time slots in the total IFF/signaling time slot. This then gives the total number of simultaneous accesses (where simultaneous is defined to be over the selected frame interval of interest; e.g., 1 second here). The result of worst case analysis is 150 for the case of no jamming and where the implementation of power control is not factored in.

The AJ capability of the IFF function involves the same processing gain of 100 MHz total bandwidth hopped over to 2400 Hz baseband, or 46 db. However, since the IFF transmission need only be 20 milliseconds long each 1 second, as described previously, one can enhance total IFF AJ by 17 db if a peak-average power trade can

be obtained corresponding to the low IFF duty cycle transmission. The time slot assigned for IFF would be permuted on a 1 second frame-to-frame basis to prevent an intelligent jammer from learning the slot. The total effective IFF processing gain then is 63 db. The same DPSK modulation/double coding combination would also be used as for COM. The fact that increased AJ processing gain is obtained for the IFF function is significant, since a non-coordinated waveform multiple access design is used, and in the COM function a coordinated waveform multiple access design is used, the actual J/S allowed is affected by the random access mutual interference present. The AJ-multiple access performance for the IFF function is indicated in Figure 4-3.

Superimposed on the figure is the AJ performance in the COM function for which the same AJ is realized up to full multiple access capacity. For IFF, it is seen that the AJ performance crosses the COM one for an IFF access capacity of 147 (this result was simply and approximately computed by adding the random access error rate to the jam error rate, when each is computed separately and independently, to get the total

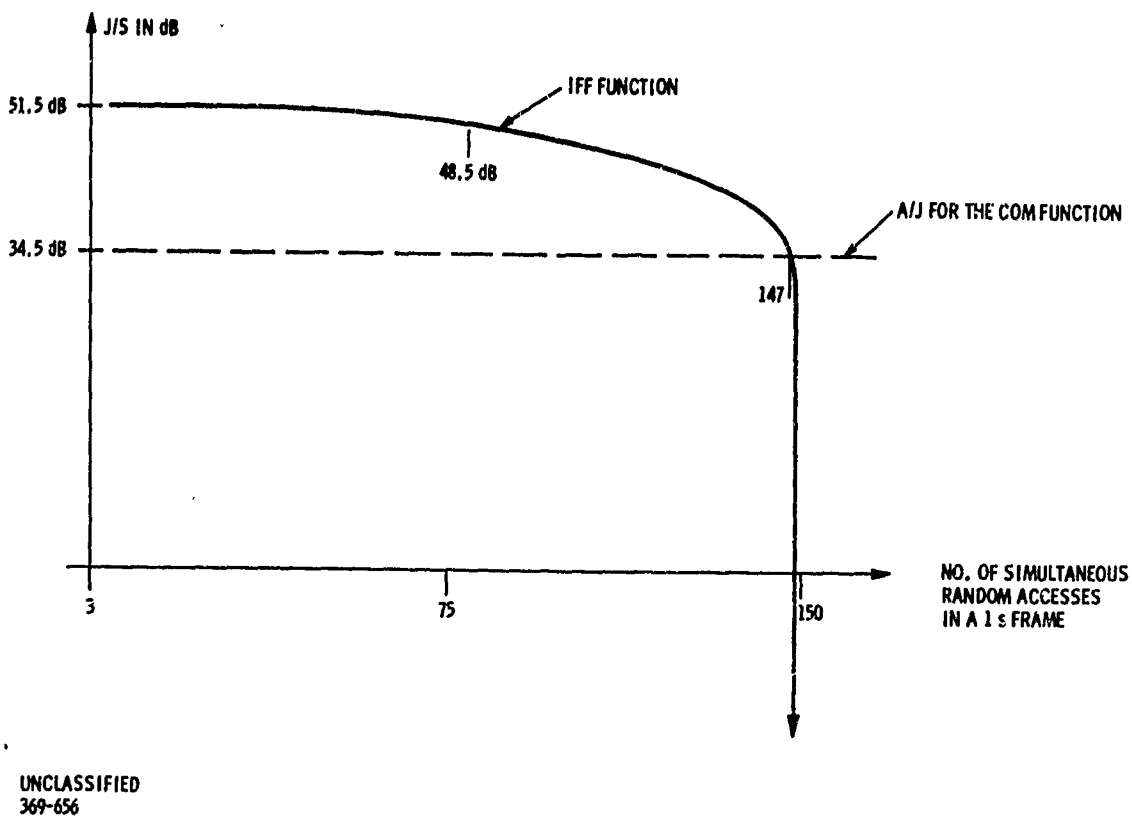

Figure 4-3. AJ Versus Multiple Access Performance for IFF Function.

channel error rate after DPSK demodulation). Therefore, at full desired IFF capacity (100), greater jam margin is obtained than for the COM function. If the peak-to-average power trade were not realized then at full IFF capacity the jam performance would be less than for the COM function.

4.1.1.3 Navigation

For the NAV function range and range rate measurements are made with clock phase and carrier track loops on the frequency-hopped signal. A coarse range measurement results from measuring the clock phase of the channel PN signal (for measurement to 1/100 of the clock phase an accuracy of 50 feet is obtained in the presence of additive Gaussian noise). A more precise range measurement is obtained by measuring the phase of a frequency hopped side tone modulation on the microwave carrier (for a 5 MHz tone, an accuracy of a few feet is obtained). Coherent frequency hop demodulation is required and instrumented in the latter case, as shown in Figure 4-4. Because of diversity action multipath is also discriminated against by coherent frequency hopping so that range accuracy in the order of a few feet is also obtained even in the presence of multipath. Now for the coarse measurement using the PN signal and with just Gaussian noise, the accuracy of 50 feet can be achieved without coherent frequency hopping. However, in the presence of multipath, an error of up to 1 mile (full clock phase) could result in this case. By making the frequency hopping and dehopping coherent, in the coarse PN measurement as well as the precise mode, diversity action can be obtained against multipath when it overlaps the PN correlation interval. A coarse range accuracy approaching 50 feet can, thus, still be obtained whether multipath is present or not.

Though sidetone ranging is indicated above as the means of obtaining precise range (reason is maximum commonality of the channelized waveform concept for NAV as for COM), an alternative would be to use an FH/PN waveform with a wider PN channel bandwidth (e.g., 1 to 10 MHz) for the NAV slot alone. The COM function would, of course, still use FH/PN with a 200 KHz PN channel bandwidth. Precise NAV range measurement would then be obtained by a "one-shot" clock phase measurement on the wideband PN signal. Tradeoffs of these two alternatives involve consideration of relative ease of implementation (such as for the wideband PN approach which does not require coherent frequency hopping).

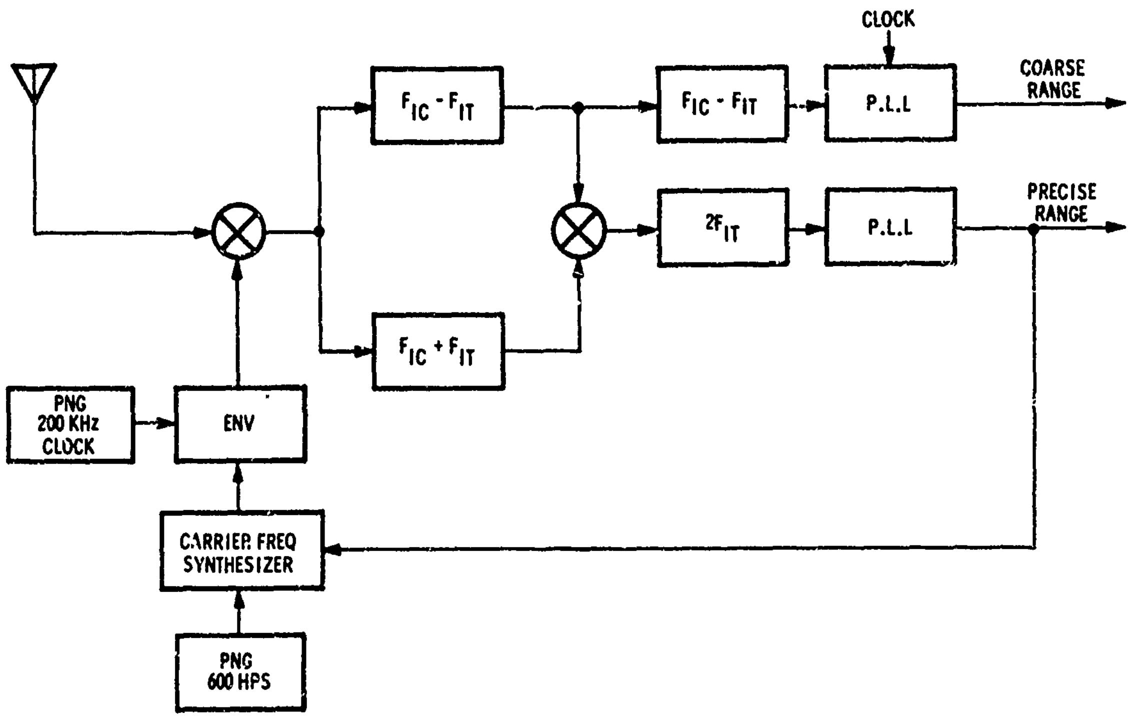

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Figure 4-4. Coherent Frequency Hop Demodulator Block Diagram.

Since the COM, NAV, and IFF functions are time division multiplexed in the concept described here, the NAV range and range rate equipment can be used for additional purposes during the IFF function time slot. Specifically whenever an IFF query is sent out, or when a random access signal is sent out so as to initiate getting into the orthogonal COM net, the NAV equipment can be used to get a measurement of the round trip propagation delay between transmission of the query or signal and receipt of a response. The fact that digital matched filters are used for sync acquisition and thus real-time acquisition is allowed, makes this transponder ranging concept feasible (the only processing delay in the transponder is now just the message demodulation time which is known apriori and therefore is compensated out). The purpose of these transponder range measurements is: (1) to give a backup range measurement, and in any case a direct range measurement to terminals of interest for reasons, say, of collision avoidance or just navigation; (2) to provide authentication of IFF and signalling replies for terminals without a key generator - one correlates the measured range with that indicated by other sensors.

4.1.1.4 Synchronization

Digital matched filter sync of the PN waveform in a frequency channel is the preferred method in the FH/PN system concept here. (See Section 7.5.2 of Volume II for further detailed performance analysis.) The NAV channel output could also be superimposed to control tap change, such as for the random IFF function. The storage requirements for the DMF are 84 bits, corresponding to a DMF processing gain of 19 db (this requirement results from the fact that to get a total processing gain of 46 db, with 27 db due from hopping over 500 frequency channels, gives, therefore, 19 db needed for the DMF - this corresponds to a 200 Kz PN channel waveform and a 2400 Hz output). The DMF receiver will actually comprise a pair of tapped delay lines each of 84 bits that are fed by the two components of the received PN signal, plus a pair of tapped delay lines each of 84 bits for the reference sequences. The 84-bit received and the 84-bit reference sequences are combined and examined for threshold exceedance. The practical implementation of a DMF will entail some processing gain loss that would have to be made up by increased ideal DMF processing gain, and therefore, greater bit storage. For example, if a conventional digital, baseband implementation of the tapped delay lines were used, the preceding amplitude and time quantization required on the received signal could entail a loss in the order of 6 to 7 db depending on the nature of the interference. A digital delay line storage of 420 (=5 x 84) bits would be needed to provide compensating processing gain.

Alternately, more sophisticated design techniques, such as multiple level quantization and bit encoding, can reduce the quantization loss and therefore the amount of compensating increase in DMF storage capacity required, but at the expense of more digital delay lines and summing logic. For the ideal 84-bit delay line configuration, it takes one output interval of 416 microseconds (for a 2400 bps output stream) for a threshold exceedance to occur out of the DMF acquisition circuit. Since a confirmation test must follow the primary one in which there is a threshold exceedance, total acquisition in this ideal case actually takes two output intervals, or 833 microseconds. For the other situation where longer DMF delay lines are used to overcome implementation losses, then acquisition would correspondingly be accomplished in up to 5 times this or 10 output 2400 bps intervals. All this presupposes frequency hopping does not occur in a DMF processing interval, for all the chips would not otherwise be correlated coherently. This would constrain the hop rate to (2400/5 = 480 hps) for the lossly implementation case, whereas for the ideal lossless implementation case,

the hop rate could be 2400 hps. The higher rate is of course desired for the random access IFF and signaling modes in particular. The above results also presume that no search is required over frequency uncertainty intervals, but just over time uncertainties.

The coordinated FH/PN waveform design, requires in addition to the link waveform synchronization time, time to get organized or coordinated among the different accesses. This is of course a penalty one pays for the advantages obtained with such waveforms over the non-coordinated. The FH system could take 30 to 50 milliseconds to organize (the time computation is based on signaling in the first of the 3 sub-time slots for the IFF and signaling function and getting a response immediately in the one or two succeeding sub-time slots). However, if a multiple signaling try is made, then waveform coordination will not occur for one second, which is the IFF signaling frame time (it was picked arbitrarily, but also to be consistent with the NAV function, for which a one per second fix rate is desired). This is still a reasonable response time. Furthermore, once organized, the FH system requires information storage only over the 50-millisecond IFF function slot, meaning therefore, only 120 bits at 2400 bps.

4.1.1.5 Multiplexing of COM, NAV, and Random IFF/Signaling Functions

Time division multiplexing of the COM, NAV and IFF functions is used in the coordinated FH/PN waveform concept illustrated here. However, the NAVSAT function can be frequency division multiplexed with the rest of CNI if the satellite antenna capability (discussed subsequently) cannot be implemented. Reference Figure 4-5 for a diagram of the TDMA operations. The NAV and IFF time slots and sub-time slots are as discussed previously. Typically, the smallest time slot would be 10 milliseconds. In the direct mode, where propagation delay differences of up to 1.8 milliseconds for 300 miles could result, the time slot width is seen to be adequate in that the required guard time to counter mutual interference is a small fraction of it. For the direct mode, this concept of TDMA functions is, therefore, seen to be reasonable (the power budget for the direct mode, though not discussed here, is also adequate), particularly if passive hyperbolic ranging is used for NAV in the terminal area and for ILS. For a beam-rider ILS scheme, the time slot concept would either require that the ILS beam be scanned rapidly in a short slot time or else that ILS be done on separate frequencies in parallel with the COM and IFF functions. The latter, of course, would represent an exception to the basic TDMA function concept and would reduce the advantages intended to be realized by it. Now for the satellite mode, in which passive hyperbolic ranging

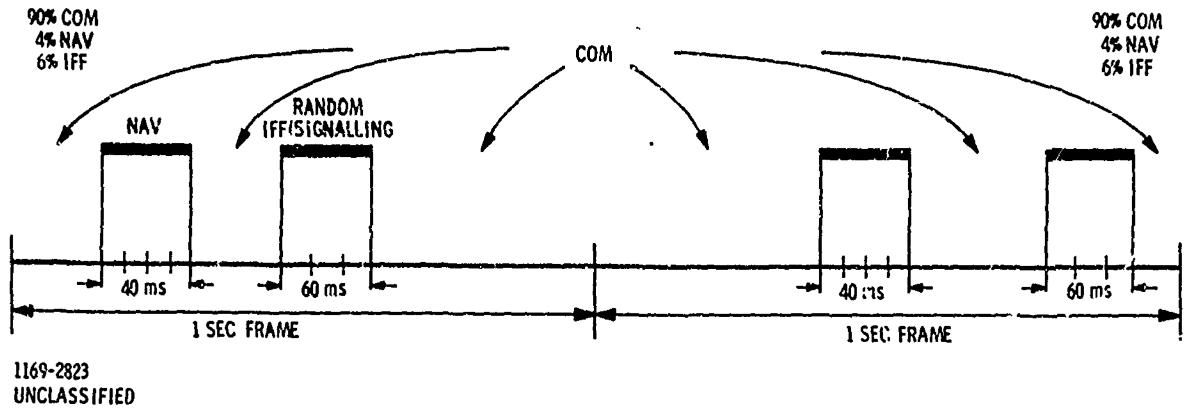

Figure 4-5. Timing Sequence of CNI Operations.

would be done for NAV, TDMA functions would, of course, also be used but special constraints must be imposed on the satellite to make it work. This is because of the power budget problem, because of the inter-satellite range differences, and because omni antennas on the aircraft make all satellites visible simultaneously. As a specific concept, steerable narrow beams are presumed on the satellite (e.g., 2^{o}). Each NAVSAT in a 621B geometrical configuration will sweep (programmed manner) its coverage area with a 2 degree beam, covering the entire area once in each one second frame interval, and dwelling on each region that is subtended by the 2 degree beam for 10 milliseconds. The NAV beam sweep for each of the 4 NAVSAT's is coordinated so that no two NAVSAT's illuminate the same region at the same time. Now other satellite beams, of about 2 degrees in width, will also be used simultaneously (i.e., in parallel) by one or more of the satellites for the COM function. That is, the power budget problem at microwave frequencies for an omni-antenna aircraft terminal requires that something like a narrow beam satellite antenna be used for the COM function. Since the concept here is to TDMA the COM, NAV, and IFF functions, the satellites must be synchronized so that when a NAVSAT 2 degree beam is illuminating a given region on earth, a COMSAT 2 degree beam that may be illuminating that same region is temporarily switched off for the duration of the 10 millisecond NAVSAT sweep interval. Because of the inter-satellite range differences to a given region on earth, appropriate compensation in satellite antenna switching time is also provided in the satellite synchronization scheme. As such, there will be no overlap between the NAV and COM functions. Similar satellite antenna beam control would be provided so as to achieve a separate IFF/signalling set of time slots. The net result is that the same timing

sequence of CNI operations as indicated in Figure 4-5 for the direct mode, will also work for the satellite mode. The narrow satellite antenna beams, besides allowing this time slotting, also provide the necessary power budget for satellite links that are power-starved because omni-antennas are used on-board the user aircraft. On the other hand, if the NAVSAT antenna design difficulties are limiting, the NAV function can be multiplexed on an FDMA rather than the TDMA basis in Figure 4-5.

Now the reasons for designing a TDMA sequence of operations for the coordinated FH/PN waveform concept here are as follows: (a) mutual interference between the 500 COM accesses and the NAV signals is eliminated - this is of particular significance to precise NAV ranging since the coding options available to protect against this interference are more limited for NAV than for COM - in any case increased efficiency is inherent in the use of coordinated waveforms; (b) time-sharing of equipments is inherently allowed - considering that one must have up to 4 NAV station fixes plus 3 IFF queries plus COM, therefore a reduction from 8 RF receivers would be required if processing were all done in parallel, to just one time-shared receiver - also other possible equipment savings result, for example, by sharing the key generator device for COM with that used to provide IFF authentication; (c) no sampling loss in S/N occurs in the process of time-sharing of equipments, such as would occur if time-sharing were applied to non-coordinated waveforms - the sampling loss corresponds to the duty cycle of receiver look at a given function - if, for example, all the 4 NAV signal sources, the 3 potential IFF queries, and COM transmissions were equally and sequentially sampled, then a 9 db sampling loss would occur to the COM channel; (d) the detection of the NAV and IFF signals in short bursts means that though these signals are of low information content the receiver bandwidths would be similarly wide as for the COM signal; i.e., 2400 Hz - the corresponding track loops should therefore hold lock for aircraft accelerations greater than 10 g's - no external aids, such as inertial platforms, are required so that the number of aircraft total black boxes devoted to the CNI function is minimized.

4.1.2 ORTHOGONAL TIME HOPPED TH/PN WAVEFORM SYSTEM APPROACH

An alternate coordinated waveform approach comprises synchronized time hopping with the accesses, as well as the functions (COM, NAV, IFF) time division multiplexed. The rationale behind consideration of such an approach is that besides having the multiple access and antijam efficiency of orthogonal waveforms, it also

does not have the power control problem among individual users nor does it have the problem of interaction among users in a nonlinear channel, since by definition only one user transmits at a time. The significance of the latter is seen in the satellite mode where for the FH/PN approach, suppression of one signal by another would occur in a hard limiter, unless effective power control were made, and also intermodulation interference results regardless of power control. The significance in the direct mode is seen for FH/PN where the hopping rate was constrained to 150 hops per second so that overlap in frequency (because of variable propagation delay) of simultaneous strong and weak signals will not occur.

Figure 4-6 indicates the signal structure of an orthogonal time hopping approach (TH/PN). Each access transmits in a separate time slot whose size is determined by the fact that in it must be included a guard time corresponding to the maximum relative propagation delay among user terminals. Designing for a 25 percent stacking efficiency (same as in FH/PN) gives a 6 millisecond slot in which a burst transmission of 4.5 milliseconds occurs. Considering 500 accesses to be in one TDMA format gives a duty cycle of 0.15 percent and a frame time of 3 seconds. The frame time can be reduced below the latter value if it is deemed to be intolerably long by designing for a smaller group of TDMA accesses, each on a different carrier, with the same total multiple access. For the long frame, a buffer storage for 7200 bits is required (obviously a problem but one that is less significant as more compact LSI storages become available). The same combination of DPSK with double coding (2 out of 3 repetitions, and 23, 12 Golay) is used for each bit, so that the number of digits required is expanded by 6 (for better performance then this one would use DPSK with just rate-1/2 sequential decoding. As indicated in Figure 4-6, the digit width needed is 0.104 microsecond (needing a 10 MHz bandwidth). To achieve a 100 MHz bandwidth so as to have 46 db processing gain, each digit is represented by 10 chips from a long PN code (quadriphase PN modulation is actually used). The total processing gain is realized by the combination of 10 db due to PN processing, 28.5 db due to the peak-to-average power trade possible at a duty cycle of 0.15 percent, and 7.5 db due to the redundancy of coding. The position of each user terminal's time slot position in a frame is pseudo-randomly permuted from frame to frame to prevent spoofing and to force a jammer to use continuous jamming.

The IFF and NAV functions are time division multiplexed with the COM signal just described, and in the same manner illustrated in Figure 4-5 (30 millisecond slots

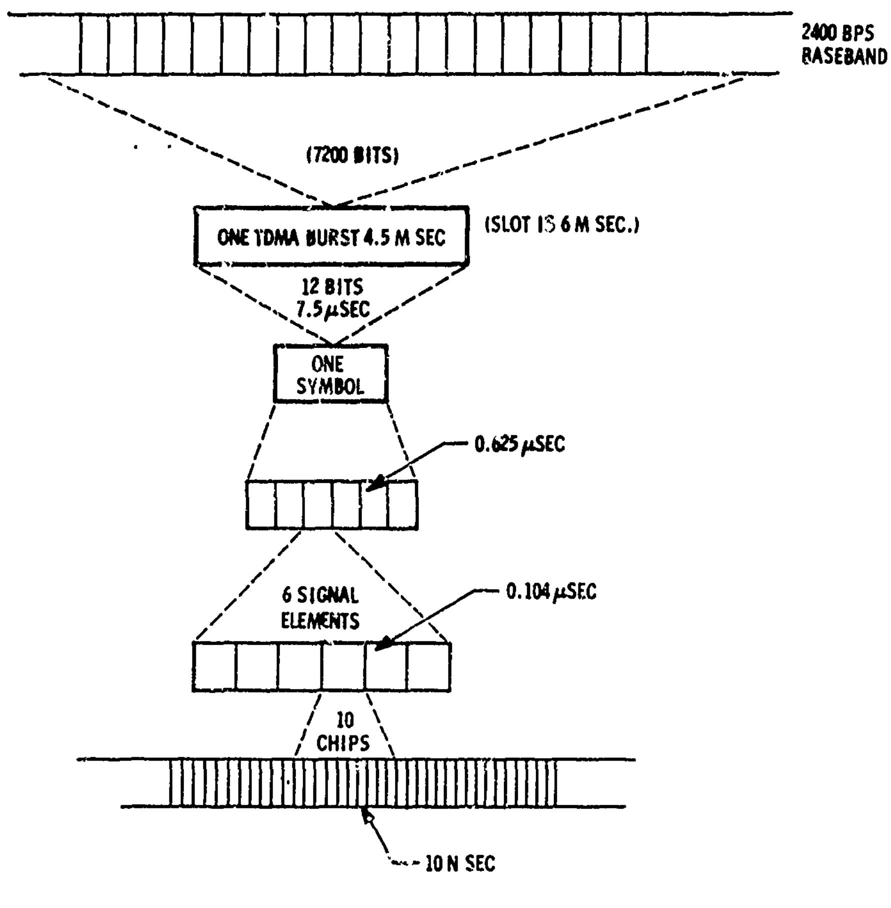

Figure 4-6. TH/PN Signal Structure (Orthogonal System).

each, every second). For IFF, asynchronous time division multiplexing is used instead. However, the same DPSK modulation and double coding combination is used and the same digit width of about 0.1 microsecond. The digit positions are permuted pseudorandomly on a frame to frame basis. The permutation pattern is also controlled by the NAV channel position fix output, so that a position address is available for each aircraft. One hundred simultaneous 50-bit IFF accesses are allowed. The reply and response to three simultaneous queries is handled sequentially in the 30 millisecond IFF function interval, as described for FH/PN.

For NAV, the time of arrival of the 0.1 microsecond pulses that are PN modulated to 100 MHz bandwidth is what is measured for range measurement. Resolution to a few nanoseconds is achievable. Range is measured to each of the reference stations, as described previously.

Multipath that arrives in a frame interval of 0.416 millisecond (about 80 miles) for NAV but after the transmitted pulse of 0.1 microsecond (100 feet) is time gated out in the receiver (only the first received pulse is accepted for range measurement). Multipath that overlaps the direct path pulse is discriminated against by the PN correlation processing. Similarly for the COM function, where a burst of successive pulses is transmitted, 10 db discrimination by PN processing will reduce the multipath intersymbol interference so that the channel modulation and coding can handle it (slightly greater than 3 db E/N_0 into the DPSK detector gives 10^{-5} bit error rate out after decoding). What this all means is that multipath interference can be effectively handled in this TH system.

Synchronization of this TH waveform in 20 milliseconds obviously is more difficult (in terms of complexity) than the FH/PN and FH approaches discussed in this section. However, it is conceptually feasible. A straightforward approach is to use a single digital matched filter, or an equivalent optical correlator, scheme. The large storage capacity required by these implementations implies that their realization depends on advances in the state of the art.

The TH scheme is not vulnerable to look-through jamming as are the relatively slow FH approaches. It has the same processing gain. Using the same coding, it has the same AJ performance for noise jamming as does the FH/PN approach described. In fact, the curves of Figure 4-3 on AJ versus number of multiple accesses also apply for TH in the direct mode for both the orthogonal and nonorthogonal waveform cases. Combination with sequential decoding would give a J/S ratio of 38 db (optimizing the modulation improves it to 41 db). However, for worst case jam strategy, the TH performance is expected to be better than for the FH/TH approach. In the satellite mode, since there are no power control problems, the multiple access performance is not expected to be degraded as would be the case for FH/PN. The jam performance in the satellite mode is still degraded by 6 db because of limiter suppression by a strong, constant envelope jammer.

Aspects of satellite interface are similar to those already discussed for FH/PN except in the area of satellite processing. In this regard, it is noted that the TH approach does not require special processing to optimize performance (except for inclusion of means to reduce the jamming limiter suppression loss). As such, the TH waveform is essentially optimum with the hard limited, straight through repeater satellites. Not including special satellite processing simplifies the cost consideration of future satellites. Also, for futuristic satellites employing a large number of extremely narrow, steerable, antenna beams (so as to enhance up-link AJ for small terminal users), the use of synchronous TH is compatible with time shared switching and use of the antenna beams.

A civilian version of the TDMA approach would be allocated a fixed time slot for COM. There would be no time slot permutation or PN spreading. Just 0.1 microsecond digits are used. It would take longer to sync than 20 milliseconds and use less costly techniques. For IFF, control of the pulse position or pulse coding, in accordance with the NAV fix would be required. The number of allowed accesses would be limited. The deletion of functions and operations of the civilian version is otherwise as described for the PN approach. However, it is noted that with regard to coexistence with existing civilian equipment, there is a spectrum allocation problem. The TH approach requires one contiguous bandwidth.

4.1.3 UNCOORDINATED TH/FH

Consideration of a uncoordinated waveform concept is made to determine the possibilities of a system that does not have most of the coordination complexities that a coordinated waveform has (in particular, the need for coordination signaling and control is eliminated, although stable clock timing is retained; also, limitations on parameters such as hop rate are relaxed). Because the need for coordination is eliminated one would also expect sync time to be minimized. However, no significant improvement should be expected in other performance areas. On the contrary, one would expect system bandwidth and power efficiency to be less for a uncoordinated waveform design.

The illustrated uncoordinated waveform considered here is a time-hopped one. Before describing it, the point should be noted that another uncoordinated, time-hopped waveform was also considered but discarded; namely, the concept originally advanced in 1967 by the CNI study group assembled at RADC. It uses a time-gated PN waveform

that is time shifted in a 2-millisecond frame to give an M'ary data modulation with M = 32. The waveform has a 100 MHz PN code rate and the receiver employs a digital matched filter with threshold detection to indicate the time of occurrence. This waveform was discarded for the following reasons:

1. Multipath:

The time-gated PN concept uses 10 nanosecond chips that are pseudorandomly gated on in an M'ary time-shift-keyed slot of 2 milliseconds interval (1000 chips on an average are on out of 6250). Since the waveforms have uncoordinated access, the system is vulnerable to multipath ringing for most multipath situations with an omnidirectional user antenna. (This is the backscatter multipath that has a level within the dynamic range limits.) That is to say, the multipath reflections for strong interferers will fill in the gated-off positions. Time-gating within a burst is ineffective, therefore. One has left then only the 32 to 1 time-gating due to M'ary time shift keying. The system multiple access capacity, thus, is reduced by the amount of the change in actual interfering duty factor, i.e., over 6 to 1.

2. Equipment Timesharing:

Since the time-gated PN receiver must sequentially check each TSK position for detection of one COM or data signal then on the average it will be tied up for 50% of the time for this purpose and would be unavailable for sharing to detect another simultaneous signal.

3. Power Loss on Satellite Downlink:

Since the time-gated PN waveform is non-coordinated, then an average power loss in use of a satellite repeater is incurred (compared to a uncoordinated waveform) because of statistical overlap of signals. Without even considering multipath "fill-in", previous analysis* has shown this to be about 7 db. If one considers, furthermore, that the M'ary PN scheme uses a digital matched filter for data detection, then the inherent implementation losses of it must be added to the total

* Aein and Schwartz, eds., "Multiple Access to a Communication Satellite with a Hard-Limiting Repeater," IDA Report R-108, Vol. II, April 1965.

satellite downlink power budget. This implementation d. m. f. loss will be 3 to 7 db or even greater, depending on the specific implementation. The total link loss then is 10 to 14 db for the time-gated PN scheme as compared to a coordinated waveform design. Since the satellite is expected to be power-starved, these degradation factors are significant.

4. Digital Matched Filter Sync:

The time-gated PN scheme requires 6250 bits of storage and a 100 MHz clock rate. This should make the design of the time-gated PN scheme complicated and costly, particularly considering implementations to reduce the loss of processing gain due to quantization.

5. Analog Versus Digital:

The time-gated PN scheme depends on analog processing to a great extent (PN correlation) for system error control, in contrast to using error correction coding in a more digital design. Overall, the latter implementation should be more desirable.

Let us now return to the time-hopped waveform that is briefly considered here. It has a time duty factor of 1/200, which also means it has good dynamic range. Furthermore, this value is presumed to be within practical feasibility in peak-to-average power trade. Also in the remote (satellite) mode it corresponds to the situation where there is plausible power available per COM signal access for peak-power limited transmitters. The waveform is also frequency hopped over 100 frequency channels, each of 1 MHz bandwidth, to cover a total of 100 MHz (the frequency channeling allows additional orthogonal slots over that possible with just time hopping, facilitates sync, and has moderate dynamic range capabilities against spectrum splatter). This gives then a total of 20,000 orthogonal slots to handle the multiple access signals. DPSK modulation together with sequential decoding to process the desired signal in the presence of multiple access interference is used to get a 10^{-5} bit error rate and reasonable AJ. An ideal maximum of 2000 strong signal accesses are possible with this system as seen from Figure 4-7, if multipath can be ignored.

This leaves, however, the effects of multipath to be considered as causing a significant interference situation. As also seen from Figure 4-7, if the multipath is of sufficiently long delay and of appropriate strength so that it contributes to multiple

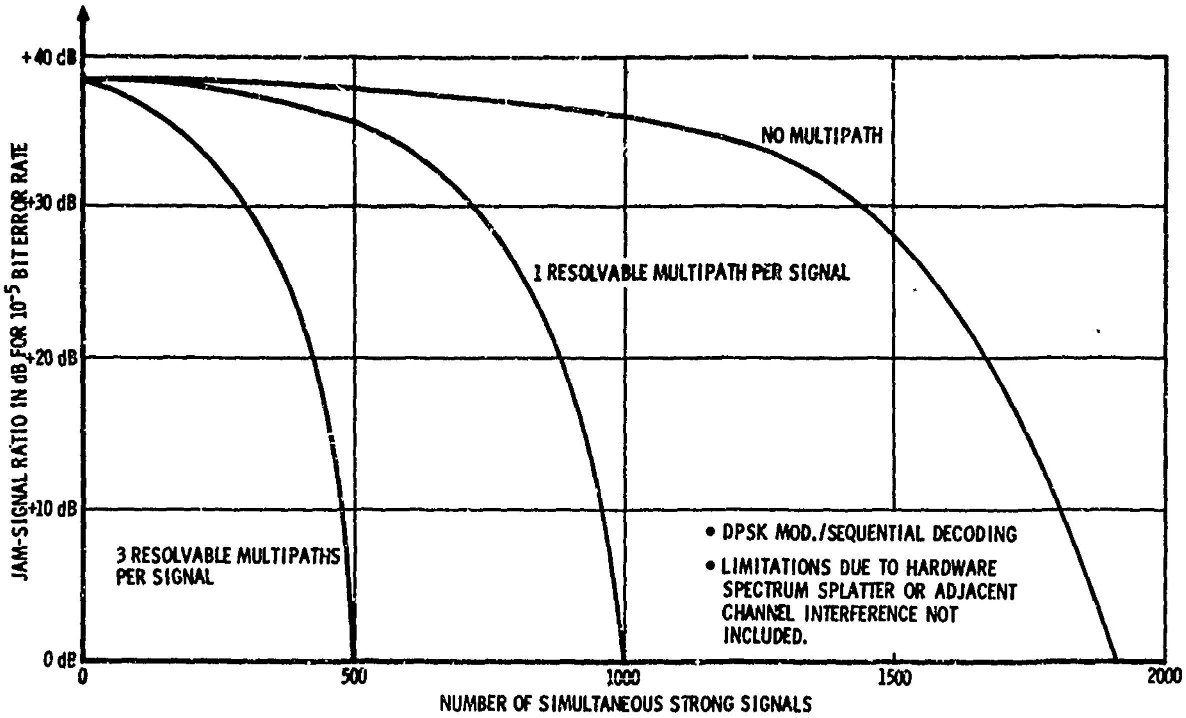

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Figure 4-7. TH/FH Uncoordinated System - AJ-Multiple Access Performance for COM for all Strong Signals.

access pulse density (this would be for backscatter multipath), the system capacity can be significantly degraded, depending on the number of reflections. Because of this, the emissions for the uncoordinated waveform are in bursts of pulses with frequency hopping on a burst-to-burst basis rather than pulse-pulse (for the above frequency-time slotting the pulses are 1 microseconds wide each and of $\cos^2 x$ shape in the time domain). The burst length is 30 to 50 microseconds so as to span the ringing time of most reflections in the system (actually in the worst case, the ringing for a high altitude aircraft would be 250 microseconds; however, to avoid buffering problems for the worst but rare case, a smaller burst length is indicated and error correction coding counted on to correct out the few long resolvable reflections occurring as multiple access interference). But now, the multipath problem becomes that of intersymbol interference among pulses in a burst. Figure 4-2, previously, indicates that with no frequency hopping in the pulse burst, a pulse signal-to-multipath interference of 3 db or greater should exist for error correction coding to be effective. This waveform is, thus, seen to be limited to multipath situations that are down 3 db or greater. The multipath impact on capacity with the pulse-to-pulse hop design as in

Figure 4-7 is now transformed to an impact on E/N_0 or error rate for the burst-to-burst hop design such as shown by Figure 4-2. The capacity for the latter system would still be high, i.e., up to 2000 signal accesses.

As an alternative to this system design one could use 10 db of PN modulation of each pulse, which would now be 10 microseconds long. The burst length would also be 10 times longer to accommodate the same information content so that the time duty factor is decreased to 1/20. For strong multiple access interferers, the 10 db PN processing is ineffective, so that system multiple access capacity is reduced from 2000 to 200 or less (if one had kept the 1 microsecond pulse width and added 10 db of PN spreading, then the number of frequency channels is reduced; the impact on multiple access is still about a 10 to 1 reduction since the number of total orthogonal slots is reduced by 10). One sees then that with the problem of intersymbol interference due to multipath solved by this waveform processing, multiple access capacity is degraded.

As a further alternative to the above solution, the use of other modulation combinations than DPSK with sequential decoding could provide an increase of 10 to 1 in ideal multiple access capacity, as discussed in Section 7.3.4 of Volume II. This alternative system would then meet the above difficulties. The quantity now traded off, however, is AJ.

Since many of the problems above hinge on the specific characteristics of the multipath and the statistical distribution of different operational situations involving it, the first time-hopped, frequency hopped waveform discussed above (i.e., without PN encoding) is retained for subsequent discussion here, at least for the COM function. A combination with PN modulation will be used only for the random IFF and NAV functions as discussed in subsequent paragraphs.

Now the only other effect of multipath not discussed yet is that of multipath fading which will occur for sufficiently short multipath delays such that overlap occurs of the multipath pulse with the transmitted pulse. Though for 1 microsecond pulses, this means a 1000 foot differential delay (such as when one is on the horizon in the air-satellite mode) so that it has small likelihood of occurring, no real problem exists anyway. This is seen in Figure 4-1 where error correction coding is sufficiently effective in handling multipath fading errors that only about a 1 db impact on E_b/N_0 required for a 10^{-5} bit error rate occurs even for multipath of equal strength (close to

the peak of the curve). Independent fading is presumed on a burst-to-burst basis, with appropriate interleaving used.

For random IFF, a similar signal structure is used, and the accesses are uncoordinated and sent in parallel. However, since a given receiver is to be designed to handle 3 low-bit queries at a time (e.g., 3 IFF queries of 50 bits each in a 1 second frame time), then the ideal processing gain is potentially greater than for the COM signal. Since it is desired to use a digital matched filter for IFF bit detection (to give address control with position) as well as sync, then the increase in ideal processing gain can be used to offset the implementation losses of the digital matched filter (in contrast for the COM function the digital matched filter is used only to aid sync). For this purpose, 12 db of PN modulation is used (ratio is 2400/150). The transmitted burst sequence of 30 to 50 pulses, 1 microsecond each, now are PN chips. The same time duty factor of 1/200 and number 100 of 1 MHz frequency channels are used as for COM. No change in tolerable, strong interferer, multiple access capacity results then, but the 12 db of PN processing is available for AJ to offset the digital matched filter implementation losses. Incidentally, multipath intersymbol interference is also discriminated against; but if the level of multipath is sufficiently down that the COM function is feasible, then multipath is not a factor for IFF either. The curves of Figure 4-7, though depicting COM AJ-multiple access performance, also give an approximate indication of IFF performance if one scales the points on the curve appropriately.

For the NAV function, a separate time-coordinated time slot is used for transmitting the NAV signal. (If satellite implementation difficulties preclude this, then one can multiplex on a uncoordinated basis.) All COM and IFF transmissions are switched off for this slot interval. The 1 microsecond pulses in the transmitted sequence are processed as PN to get range measurement to 10 feet at best, but practically, perhaps only to 60 feet. For more precise range measurement, use of a 10 MHz PN waveform would be a straightforward alternative except that the digital matched filter is now more complex. The use of a digital matched filter is deemed desirable, particularly for NAV, to allow rapid reacquisition at each fix interval that is a 1 second frame interval apart. The ILS function would also be handled similarly except that multiple, spaced slots in a 1 second interval would be used to get a more rapid fix rate.

The satellite mode for this concept would at least have the downlink in the same frequency band as the direct mode. These downlink transmissions will be non-coordinated and use the same signal structure as for the direct mode. As indicated in Figure 4-7, adequate, ideal, multiple access capacity exists to handle these accesses. The satellite uplink on the other hand, would be in a separate frequency band.

Refer to Table 4-1 for a summary of the key characteristics of frequency-time usage for the uncoordinated TH/FH waveform system. The numbers indicated are only for an illustrative system design. An optimized design set of parameters requires a tradeoff of the various conflicting design criteria.

Table 4-1. Summary of Frequency and Time Slot Usage for the Uncoordinated TH/FH System.

- Pulses 1 microsecond wide and in a burst of 20 to 50 of them are transmitted at a time for COM. Time duty factor is 1/200. Buffering of the pulses is used.
- Frequency hopping over 100 frequency channels (in 100 MHz bandwidth) on a burst basis.
- DPSK modulation/sequential decoding is used for COM for a redundancy per bit of 2 to 1. A burst represents then 10 to 25 bits.
- All COM signals in direct mode and satellite downlink are uncoordinated and in same frequency band. For example, 400 or more signals for former and 100 for latter.
- Satellite uplink is in a separate band with a translating repeater; this eliminates coordination with direct mode.
- IFF is non-coordinated with COM and uses same frequency band. However, a 16 to 1 PN encoding of each digit is used since information rate need be only 150 bits in 1 second (i.e., 3 separate 50 bit queries each second). IFF signal structure is same as for COM. However, a pulse burst of 30 to 50 microseconds represents PN chips for IFF information bits. There are 100 total IFF signals in same band as for COM.
- NAV (enroute, terminal, and ILS) is done in separate coordinated time slots. These are 10 milliseconds per NAV range fix. All COM and IFF transmissions are inhibited in this time. The pulse burst sequence of 1 microsecond pulses is processed as PN for range resolution. Frequency hopping of bursts of these pulses occurs over 100 MHz. However, for ranging precision a sidetone is used.

4.1.4 UNCOORDINATED FH

A generic type of system design having a number of favorable attributes and which is appropriate to the CNI problem is frequency hopping. Two variants distinguished are coherent and noncoherent frequency hopping. Coherent frequency hopping is a signal in which the instantaneous phase of the carrier is deterministic from hop to hop, whereas in noncoherent frequency hopping there is no controlled phase relationship from hop to hop. For purposes of communication, IFF, and coarse navigation, noncoherent frequency hopping suffices; but for fine navigation, the coherent version appears necessary. In this discussion, the noncoherent version is described first and then the coherent version. A distinct possibility would be to use coherent signals, ignoring the coherence for all except the high accuracy navigation receiver/signal processors.

4.1.4.1 Noncoherent Version

While the precise determination of such signal parameters as hopping rate, frequency increment, data modulation, and error coding depends on detailed analysis and tradeoffs, to make the following descriptions clearer, some plausible values have been chosen.

An appropriate hopping rate is 5000 per second, that is, the carrier frequency is selected pseudorandomly from a large number of possible equally spaced values, with a new choice every 200 microseconds. Addressing for purposes of distinguishing nets (but generally not among parties in a net) is through the use of different sequences.

4.1.4.1.1 Data Modulation

For 2400 bit per second communications and perhaps for data accompanying identification and navigation signals (authentications, altitude replies, station coordinates, etc.) if the same 2400 bit rate is appropriate, the modulation could advantageously be chosen as a 4-tone MFSK. Blocks of 12 bits would be encoded to 24 bits, using the Golay Code. Then the encoded binary symbols of six such Golay words are interleaved, the composite rate being 4800 binary symbols per second. Each successive pair of binary symbols is used to select one of the four tones. After three such pairs (and tone selections) two more pairs are calculated as parity checks, which are also used to select tones. Interlaced with this stream of frequency-shifted hopped-carrier bursts is a series of unmodulated bursts, in particular every fifth hop is without modulation. This will facilitate code tracking, resynchronization, AGC and similar operations incident to the communication function.

This timing arrangement is illustrated in Figure 4-8. In the 5 milliseconds shown, 24 data bits are accepted from the source and are pairwise used to determine the frequencies (f_A, f_B, f_C or f_D) to be transmitted during 12 of the 25 hops during that same 5-millisecond interval. From each three of the twelve frequency choices, two more choices are computed according to the coding rules described below. Finally every fifth hop is allocated to an unmodulated burst.

Superimposed upon this information modulation will be the carrier hopping, so that each of the frequencies selected here will be translated by a pseudorandomly selected carrier frequency, changed at 5000 times per second.

4.1.4.1.2 Data Coding

In hopping systems for antijam or multiple access purposes, it is essential that the obliteration of a large fraction of the individual hopped bursts be tolerated, thus implying the need for redundancy and the consequent ability to correct errors and/or erasures.

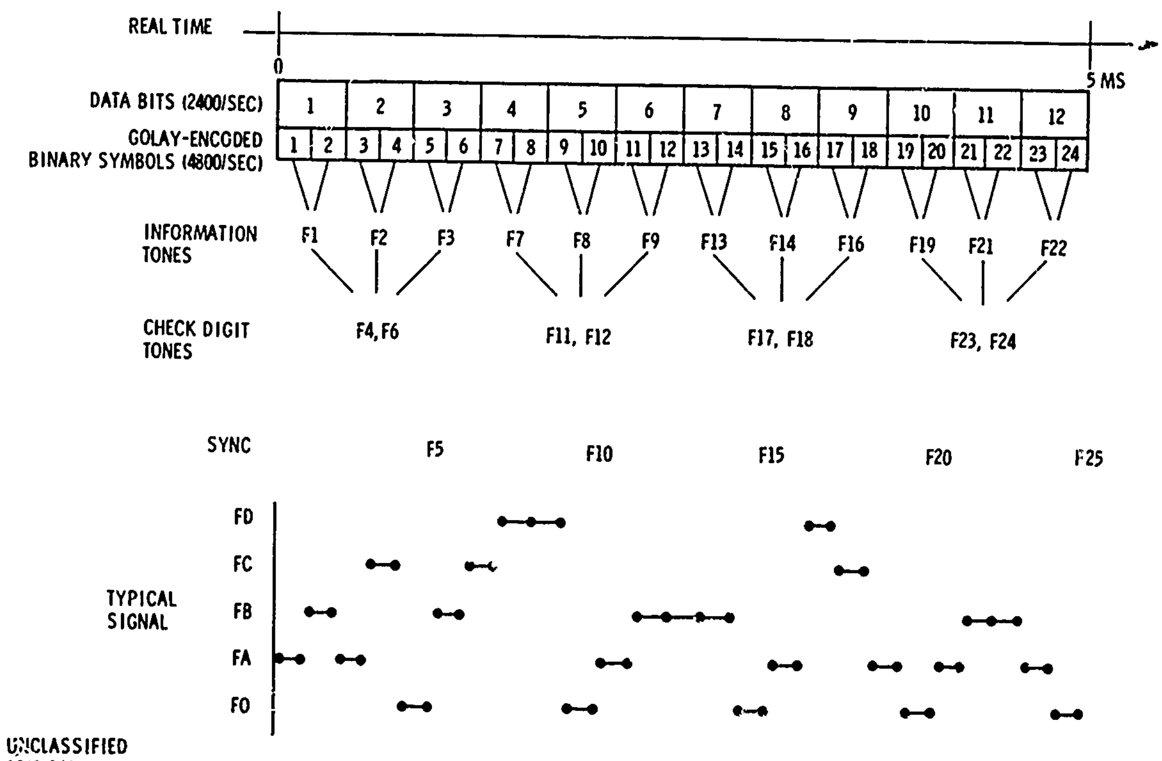

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Figure 4-8. Modulation Scheme for 2400 Bit Per Second Data.

Coding may be viewed in a different way. One may think of a frequency-hopped system as a kind of frequency diversity system used to combat a signal-to-noise ratio that varies across the band. This variation may be due either to signal variations (selective fading due to multipath, antenna patterns, propagation effects) or to noise variations (interfering signals, CW jammers, RFI). The coding is then simply a way of achieving the necessary time-frequency diversity and the decoder is the diversity combiner.

To achiev a high design of users in the allocated spectrum, extremely high resistance to mutual interference is needed, since in high density situations, the incidence of two or more signals at the same frequency will be high.

For this generic signal design, a double coding is chosen. First, the source data is encoded by the highly efficient and powerful Golay code, and then by a quaternary code, which is also very efficient. The Golay code appends 12 parity checks to a block of 12 bits. It then becomes possible to correct for as many as three errors in the 24 symbol word and to detect the occurrence of four errors. This latter feature can be useful in data transmission where it is best to discard questionable data rather than risk an erroneous decoding. In fact, using the same encoding, error correction power can be traded for error detection power at the receiver. For voice transmissions, the detection feature is of no use, but since removing it only saves one check digit, it is best retained for purposes of standardization and simplicity of timing circuitry.

The second code chosen for this candidate system is the quaternary (5, 3). Its attraction is that it is easily mechanized in a few MSI chips and that it is an ideal code in the sense of being close packed and hence efficient of energy. (This is a property also shared by the Golay code.) Since this code is not as well known as the Golay, it will be described in some detail here.

We shall use the symbols A, B, C, and D to represent the four numerals of the quaternary number system and Q1, Q2, Q3, Q4, and Q5 to represent the five quaternary digits comprising a code word. Q1, Q2, and Q3 are regarded as information bearing (six bits) while Q4 and Q5 are check digits computed from them. Figure 4-9 shows the rule for check digit computation graphically. In an algebraic representation

$$Q4 = Q1 + Q2 + Q3$$
$$Q5 = Q1 + (C \times Q2) + (D \times Q3)$$

where the operations "+" and "x" are defined as follows.

+	A	B	C	D
A	A	B	C	D
B	B	A	D	C
C	C	D	A	B
D	D	C	B	A

;

x	A	B	C	D
A	A	A	A	A
B	A	B	C	D
C	A	C	D	B
D	A	D	B	C

Study of these tables or equations will quickly convince the reader of the properties of this code, notably that the Hamming distance is 3, that it is single error correcting, that it is double erase correcting and is close-packed. Receivers could be built that employ an error correction detection scheme or a forced double-erasure scheme for this code. However, for best results, it is suspected that the proper detection scheme will be a matched filter arrangement preceded by a clipper to limit the effect of impulse interference.

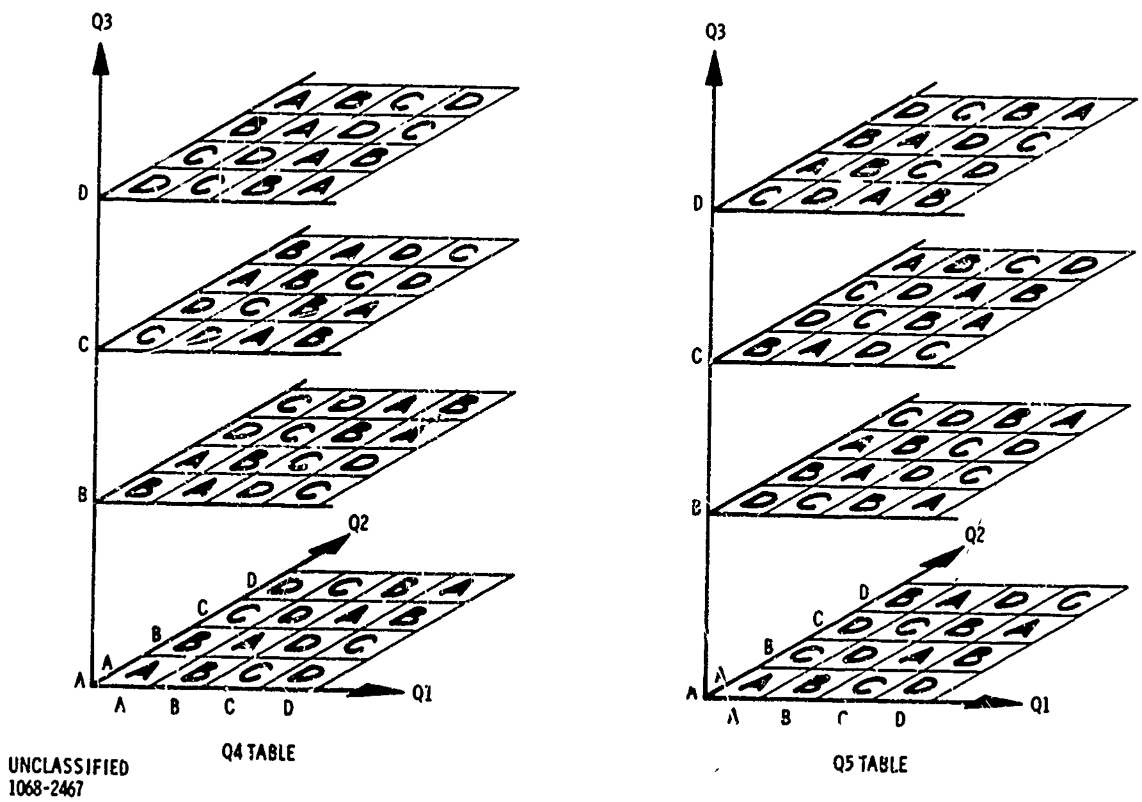

Figure 4-9. Check Digit Tables for (5, 3) Code.

A precise analysis of this scheme has not been carried though; however, indications are that the performance will be approximately equal to that of coherent binary PSK - i.e., the Gaussian noise error rate on the 6-bit block will be about 0.5×10^{-3} at 8.2 db E_b/N_o. The power of this scheme is not so much its adequate performance in impulse noise, but that it will permit one or two very strong interferences during the five symbols. This will provide immunity to occasional obliterations of pulses due to coherent strong signals.

4.1.4.1.3 Pulse Shaping

In this frequency hopping, illustrative waveform, it is desirable to use a pulse shape on the individual 200 microsecond pulses with a more compact spectral density than would be obtained with simple rectangular pulses. A good choice would be a raised cosine (or cosine squared) envelope. The spectra for these two waveforms, for equal energy pulses, is shown in Figure 7-17 of Volume II. The advantage of the raised cosine pulse becomes evident when considering the problem in the direct mode when a nearby strong transmitter is operating in the same frequency band as a distant weak transmitter whose signal is to be extracted. If the strong signal were of the rectangular type it would obliterate large segments of the band in the neighborhood of its instantaneous frequency, not just its nominal channel. However, by using these shaped pulses, the effect is greatly reduced, for example, by 30 db at channels 4/T (20 KHz) away from nominal.

4.1.4.1.4 Frequency Spacing

Having chosen the pulse shape, it is then possible to select the 4 FSK tone spacing. It is found that two such pulses are orthogonal if their carriers are spaced at any integer multiple of the pulse rate. A practical choice is 15 KHz, which would than place the four tones at -30 KHz, -15 KHz, +15 KHz, and +30 KHz from the nominal (hopped) carrier, corresponding to the four code symbols A B C D and the four tone frequencies f_A f_B f_C f_D mentioned earlier.

This gives a nominal signal bandwidth of 30 KHz, measured between the outer first nulls of f_A and f_D. If the objective of the signal design were primarily random access, then the granularity for the frequency hopped signal would be 80 KHz (or perhaps 100 KHz for compatibility with existing channel spacing standards). If jamming immunity is a serious consideration, then narrower spacing of the allowable

center frequencies would be used, possibly 10 KHz, in the interest of making the system less vulnerable to CW or similar signal jamming.

4.1.4.1.5 Spectrum Usage

A first-cut estimate of the spectral occupancy of this signal design in a random multiple access system is made here. The entire band is taken to be N times 80 KHz, with M users simultaneously transmitting, each using a "randomly" chosen channel of the 80 KHz channels. The probability that the M + 1st user will choose an unoccupied channel is then

$$\left(1 - \frac{1}{N}\right)^{M}$$

for each hop. The action of the limiter matched filter arrangement is such that all signals more than a few db below the desired signal are very sharply discriminated against and even those of about the same strength will often be sharply discriminated against when their cosine squared pulses are out of time step with that of the desired signal. On the other hand some very small number of interfering signals of very great power will have spectral sidebands which are large compared to the desired signal and effectively occupy multiple channels. For this first estimate it will be assumed that half of the total number of signals are of sufficient size to bother the detection of the desired signal. It should be noted that in certain short range communications, such as landing and tanker refueling, the detections will be much smaller. The use of directive antennas would also reduce this factor. For the typical case the probability of an effectively clear channel is thus taken as

$$\left(1 - \frac{1}{N}\right)^{M/2}$$

In attempting to detect one of the four FSK frequencies, only three out of five interference situations will result in a potential error, since one-fifth of the interfering pulses will fall at the carrier which is not being considered at the time of information pulses, and another one-fifth will be at the same frequency as the desired signal, usually reinforcing it.

Assuming a high signal to natural noise ratio, the quaternary code detection scheme described earlier, notably the clipper-matched filter, will not fail unless three or more strong bad hits occur. This is largely due to the clipping which is set by AGC action at a level just above that of the desired signal level, so that a strong hit has only

slightly more weight in the decision than a proper pulse. Accordingly, the probability that the quaternary decoding will fail is given by

$$\binom{5}{3} p^3 (1-p)^2 + \binom{5}{4} p^4 (1-p) + \binom{5}{5} p^5$$

where

$$p = \frac{3}{5}\left[1 - \left(1 - \frac{1}{N}\right)^{M/2}\right]$$

In this approximation, the occurrence of multiple interferences at a single pulse time is neglected. The quaternary decoding, when it fails, will usually give about four of the six information bits in error. This permits calculation of the bit error rate as a function of M/N, shown in Figure 4-10, assuming large N.

The Golay decoding, done on a triple error detection, then removes most of these residual errors. The straightforward calculation of its action gives the overall result, also shown in Figure 4-10.

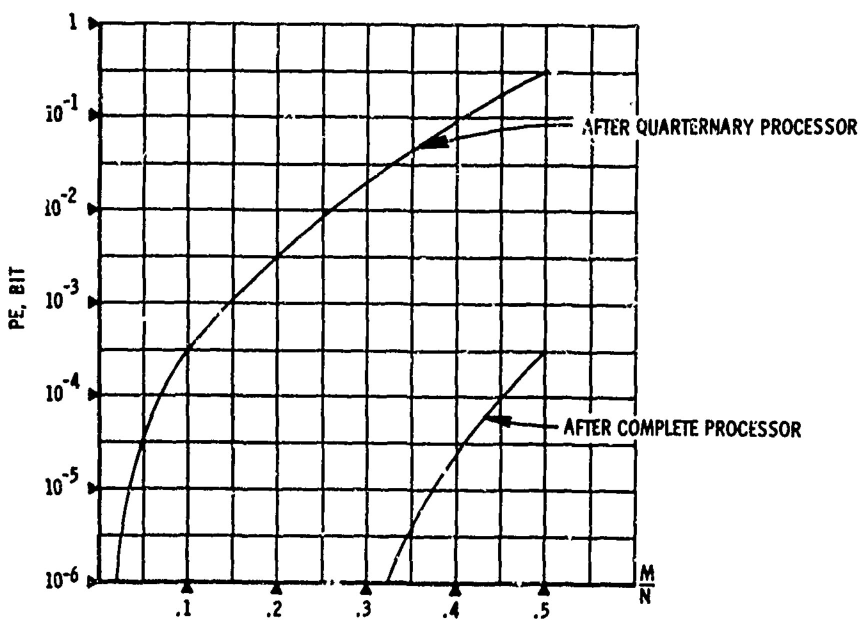

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Figure 4-10. Performance of the Detectors Versus Ratio of Users to Frequency Channels.

This estimate is that for 10^{-5} bit error rate the number of channels should be about three times the number of simultaneous "talkers." Thus for 600 channels, about 3 x 600 x 80 KHz or 144 MHz would be needed.

4.1.4.1.6 Synchronization

It would be reasonable to suppose that even the poorest user could carry a clock good to 1 millisecond absolute accuracy. Since the propagation time uncertainty is of the order of 2 milliseconds, there is little payoff in having much better clocks. The composite time uncertainty of about 3 milliseconds is then equivalent to 15 hop pulses. A simple serial search detection technique would permit each of these pulse positions to be searched in four or five information bit times, giving reliable synchronization in about 15 to 20 milliseconds. (See Section 7.4.5 of Vol. II for more detailed sync analysis.)

Sync tracking, Doppler tracking and AGC action would be performed on the interlaced, unmodulated hops, with the allocation of one-fifth of the signal energy to such unmodulated hops sufficient to ensure excellent performance in these functions.

In tracking the unmodulated pulses, it should not be difficult to interpolate between the half amplitude points on the pulse envelope (100 microseconds apart) to about 6 microseconds, given reasonable signal-to-noise ratios and smoothing times. Thus a range measurement accurate to the order of 1 mile can be very simple and quickly obtained without any elegant detection or tracking systems.

4.1.4.2 Coherent Frequency Hopping

To achieve extremely precise ranging, it is necessary to employ bandwidth much greater than that of the pulse envelope. One method which appears promising, is to make the hopping carrier coherent. As described in Section 5.2.2.2, certain frequency synthesizer configurations can be built in such a way that the output phase is deterministic. Expressed in another way, if the synthesizer were so constructed that upon selecting any multiple of the frequency standard (80 KHz in our case) the output waveform's zero crossings always coincided with those of the standards, it becomes a simple matter to measure the relative phase of two such standards driving two such synthesizers to extreme accuracy by making phase comparisons on the synthesized outputs as identical frequency selections are simultaneously altered on the two synthesizers.

Accordingly, a system is envisioned in which the navigation base station (or stations) use a very stable source and coherent synthesizer. The actual signal, as described previously, would be modulated with either data pertinent to the navigation function (base station coordinates, ephemeris, atmospheric correction factors) or otherwise. After an aircraft acquires the signal on a noncoherent basis, its navigation signal processor would then walk the phase of the frequency reference until the phase of the unmodulated (every fifth) pulses in the IF becomes consistent. A serial search, implied in this statement, would improve the 6 microsecond accuracy estimate, available within a few milliseconds of first signal acquisition, down to a few nanoseconds in about 10 seconds. If faster refinement is desired, more elegant (but expensive) methods of estimation and search could be implemented.

4.1.5 HYBRID COORDINATION WAVEFORM APPROACH

The hybrid approach illustrated here uses uncoordinated waveforms in the direct mode and coordinated waveforms in the satellite mode. Also, even variation in waveform characteristics and parameters should be considered for these two modes. The rationale for this hybrid waveform is that the operating environment (propagation differences and even threat differences) is such that no one pure waveform is optimum for both modes. A hybrid waveform design does not necessarily mean a doubling of black boxes. Instead, if the same waveform types are involved, then one black box, but with parameters and functional controls that are automatically changed from direct to satellite modes, is involved here.

An illustration of the hybrid is indicated here as a generic waveform. Specifically, a hybrid could comprise TH/FH-nonorthogonal in the direct mode, and TH/FH-orthogonal in the satellite mode. The direct mode scheme would be as described in paragraph 4.1.3. The remote mode scheme would use a 25 millisecond frame (rather than 3 seconds as in the coordinated TH scheme in paragraph 4.1.2). The burst length would be 50 microseconds. Each pulse has a 0.1 microsecond width and is hopped in frequency over 100 MHz at 19,200 hops per second on a burst-to-burst basis. There is a dehopper in the satellite so that a jammer cannot get through to cause an additive 6 db limiter suppression effect.

A further waveform possibility could be considered in the hybrid approach to use SSMA of PN waveforms in the satellite mode. That is, an SSMA approach was not considered for one of the pure generic systems because the multiple access efficiency is

low in the direct mode when there is a significant "near-far" problem. However, in the satellite mode, where effective power control can eliminate "near-far" problems, the multiple access efficiency attainable with SSMA is more than adequate. It has excellent A/J. The major problems actually are those of having a satellite waveform that is significantly dissimilar to the direct mode solution, ability of fast sync of an extra wideband PN waveform, and feasibility of such continuous wide bandwidths both in terms of waveform synthesis and the dispersion effects of the satellite channel. The latter two problems, however, appear solvable. Another factor in favor of PN in the satellite mode is that precise ranging can be obtained even under sophisticated jamming (look-through jamming is not feasible) and multipath. Direct sequence PN is currently recommended as the signal structure for NAVSAT, so that compatibility exists in this regard.

4.1.6 SUMMARY OF THE FIVE WAVEFORM TYPES

A summary of the five waveform types that have been discussed is given in Table 4-2 in terms of the key identifiable characteristics, advantages, and disadvantages when these waveforms are designed for the CNI problem here. As seen from Table 4-2, no one waveform has all the advantages stacked in its favor, and the different waveforms tend to complement each other in these regards. In particular, it appears that the coordinated waveform types have a higher communication efficiency (measured in multiple accesses, AJ, dynamic range, and tolerable thermal noise for a full capacity system) and navigational accuracy as compared to the uncoordinated waveform types. On the other hand, the coordinated waveform types required a coordination time, increased complexity because of the need for coordination, and a possible increased intelligent AJ vulnerability because of the special requirements for coordination. This is all in contrast to the case for the uncoordinated waveforms.

Now, in comparing the frequency hop type with the time hop type, whether for the coordinated or uncoordinated cases, the FH types seem to have increased communication efficiency for the direct mode, were it not for dynamic range spectrum splatter, whereas the time-hopped type has increased efficiency for the satellite mode. In the direct mode specifically, the factors of less multipath degradation, improvement through use of power control, and greater ease of sync contribute to the FH advantages. However, the direct mode efficiency of FH waveforms becomes limited in wide dynamic range situations (say greater than 80 db). Also, these waveforms have limited connectivity. In contrast, the TH waveforms have high performance in these latter regards,

Table 4-2. Summary Comparison - Generic Waveform Types
1. FH/PN Coordinated Waveform

Characteristics	Advantages	Disadvantages
▲ FH/PN coordinated for COM and NAV. ▲ FH/PN noncoordinated for IFF and signaling. ▲ TDMA modes. ▲ Sidetone or PN for NAV. ▲ DMF for sync and address control ▲ Hyperbolic NAV for enroute and landing. ▲ Steerable antenna on satellite. ▲ Processing in satellite desirable. ▲ Error correction code/binary modulation. ▲ Use common relay nodes to aid coordination. ▲ Use modes to aid each other for increased performance. ▲ Power Control.	▲ High COM and NAV AJ-MA efficiency. ▲ Further optimize AJ-MA efficiency because of coordination of modes. ▲ Time share equipment without loss of efficiency because of TDMA modes. ▲ Burst transmit of NAV and IFF gives protection against range acceleration. ▲ Effective against multipath because of long duration chip and coding. ▲ Precise NAV ranging and efficiency with sidetones. ▲ Compatible with satellite processing. ▲ Spectrum allocation compatibility. ▲ Ease of sync and address control (with DMF) and implement. ▲ Ease of synthesis of extra wideband.	▲ Limited to slow hopping in direct mode (150 hps). ▲ Coordination time and complexity for organizing waveforms. ▲ For hard-limiter repeater need power control; also intermod exists. ▲ Spectral splatter control problem in dynamic range. ▲ Depends on satellite channelized processing for mix of user terminal classes. ▲ Cannot transmit and receive on same or adjacent frequencies, except when there is antenna isolation. ▲ Coherent frequency hop synthesizer advanced implementation. ▲ Limited connectivity.

Table 4-2. Summary Comparison - Generic Waveform Types (Continued)
2. TH/PN Coordinated Waveform

Characteristics	Advantages	Disadvantages
▲ TH/PN coordination for COM and NAV. ▲ TH/PN noncoordinated for IFF and signaling. ▲ Steerable antenna beams in satellite. ▲ TDMA modes. ▲ DMF for sync and address control. ▲ Error correction code/binary modulation. ▲ Hyperbolic NAV. ▲ PN for NAV or pulse-time-of-arrival for NAV.	▲ High COM and NAV AJ-MA efficiency because of coordinated waveform and coordination of modes. ▲ Wide dynamic range. ▲ Maximum connectivity. ▲ Time share equipment without loss of efficiency because of TDMA of modes. ▲ No satellite power control or inter-mode problems; can handle mix of terminals. ▲ Compatible with serial switched antenna concept on satellite. ▲ Burst transmission of NAV and IFF protects against range acceleration.	▲ If coordination for direct pt-pt traffic then long frame times required. ▲ Requires contiguous wideband spectrum. ▲ Requires expensive DMF for high clock rate and large storage. ▲ Time dispersion problem for wide bandwidths. ▲ Coordination time and complexity for organizing waveforms. ▲ Requires high peak-to-average power trade for efficiency; a serial switched antenna beam must have high gain.

Table 4-2. Summary Comparison - Generic Waveform Types (Continued)
3. TH/FH Noncoordinated Waveform

Characteristics	Advantages	Disadvantages
▲ Modes are also noncoordinated. ▲ Pulse time-of-arrival or PN for NAV. ▲ Receiver and antenna sampling so as to time-share equipment. ▲ Either diversity combine or error correction code; also either M'ary or binary modulation. ▲ Steerable satellite antenna. ▲ DMF for sync and address control.	▲ No coordination time for accesses or modes. ▲ Achieve goal of common waveform for all modes and functions. ▲ No satellite power control or inter-mode problems. ▲ High dynamic range. ▲ Good connectivity.	▲ Inefficiency in AJ-MA because of noncoordinated accesses. ▲ Inefficiency in AJ-MA because modes are noncoordinated. ▲ Vulnerability of NAV and IFF channels to range acceleration because transmissions are at low bit rate in parallel. ▲ Degrade AJ-MA efficiency in direct mode due to multipath. ▲ Requires high peak-to-average power trade for efficiency.

Table 4-2. Summary Comparison - Generic Waveform Types (Continued)
4. FH/PN Noncoordinated Waveform

Characteristics	Advantages	Disadvantages
▲ Either pure FH or hybrid of it with some TH or PN for direct and satellite modes, uncoordinated. ▲ Modes are uncoordinated. ▲ Sidetone or PN for NAV. ▲ Processing in satellite desirable. ▲ Steerable antenna on satellite. ▲ Power control in direct and satellite mode. ▲ DMF for sync and address control. ▲ Receiver and antenna sampling so as to time-share equipment. ▲ Either diversity combining or error correction coding; also either M'ary or binary modulation.	▲ No coordination time for accesses or modes. ▲ Achieves goal of common integrated waveforms for all modes and functions. ▲ Effective against multipath because of long duration chip and processing. ▲ Precise NAV ranging and efficiency with sidetones. ▲ Compatibility with satellite processing. ▲ Spectrum allocation compatibility. ▲ Ease of sync and address control (with DMF) and ease of implement. ▲ Increased MA efficiency because of power control in direct mode.	▲ Inefficiency in AJ-MA because of noncoordinated accesses. ▲ Inefficiency in AJ-MA because modes are noncoordinated and there is a sampling loss when one time-shares the receiver and antenna. ▲ Vulnerability of NAV and IFF channels to range accelerations because transmissions are at low bit rate in parallel. ▲ For hard-limited satellite repeater, needs power control; also intermod exists. ▲ Spectral splatter control problem in dynamic range. ▲ Depends on satellite channelized processing for mix of user terminal classes. ▲ Cannot transmit and receive on same or adjacent frequencies except when there is antenna isolation. ▲ Coherent frequency hop synthesizer advanced implementations. ▲ Limited connectivity.

Table 4-2. Summary Comparison - Generic Waveform Types (Continued)
5. Hybrid TH/FH Waveform - Noncoordinated in Direct Mode and Coordinated in Remote Mode

Characteristics	Advantages	Disadvantages
▲ Accesses in each mode are uncoordinated for direct but coordinated for satellite. ▲ TDMA of COM, NAV, and IFF mode modes for direct and satellite. ▲ Steerable antenna beams on satellite. ▲ DMF for sync and address control. ▲ Error correction code/binary modulation.	▲ Reasonable direct mode design results with a waveform which when coordinated in the satellite mode is highly efficient. ▲ Time share of equipment without loss of efficiency because TDMA of modes. ▲ Burst transmission of NAV and IFF protects against range acceleration. ▲ No satellite power control or intermod probability; can handle mix of terminals. ▲ Compatible with serial switched antenna concept. ▲ Can transmit and receive on same frequency (but needs twice the bandwidth). ▲ Wide dynamic range. ▲ High connectivity.	▲ Reduced AJ-MA efficiency for noncoordinated accesses in direct mode. ▲ Degrade AJ-MA efficiency in direct mode due to multipath. ▲ Requires contiguous wideband spectrum. ▲ Requires expensive DMF for high clock rate and large storage. ▲ Requires high peak-to-average power trade for efficiency; a serial switched antenna beam must have high gain. ▲ Coordination time and complexity for organizing waveforms in satellite mode.

whereas for moderate dynamic range situations they may not be as good. Now in the satellite mode specifically, the factors of not needing power control, and of not having intermod problems and of not needing to back off in power efficiency, contribute to the TH advantages over the FH waveforms.

The results developed above would indicate that the candidate waveform designs to be selected for further study and comparison would be hybrid variations of the above basic types. Doing so would combine the best features and minimize the disadvantages of the individual waveform types, since as already stated, no one of the waveform types considered has all the advantages stacked in its favor.

4.2 CANDIDATE WAVEFORMS

As a result of the discussion in Section 4.1 on the illustrated generic waveform types, three candidate waveform designs are defined here that are in each case actually different hybrids of the generic waveform types. The waveforms are identified as FH/PN/TH-coordinated; TH/PN/FH-hybrid-coordinated; and FH/TH-hybrid-coordinated, respectively. These candidate waveforms are described now in more detail, and their rationale also indicated so that a selection of a single plausible CNI waveform design can be made in Section 4.3.

4.2.1 FH/PN/TH-COORDINATED WAVEFORM

The FH/PN/TH-coordinated waveform candidate results essentially by time-gating on a TDMA basis the FH/PN waveform design discussed previously, scaling the parameters accordingly, and also optimizing the design. The various performance factors that motivate this candidate waveform design are:

1. Coordination of the COM waveforms among the different users allows efficient usage of satellite power (main limitation on capacity in satellite mode) and direct mode bandwidth (since power is not a problem here) so as to achieve a high capacity and efficiency in all cases. The use of common satellite timing facilitates this coordination for both the satellite and direct modes. In fact it appears plausible to be able to coordinate not only the direct mode COM accesses among themselves and satellite COMSAT accesses among themselves, but also to coordinate the direct mode with the satellite downlink so that the two can use a common frequency band.

Though the NAV and IFF waveforms are coordinated in a more limited sense, or not at all, as will be indicated in the subsequent system description, (actually the NAVSAT's are not coordinated with the COMSAT; also the IFF accesses in the direct mode are not coordinated among each other) this does not really impact on satellite power or direct mode bandwidths. Reference to this waveform as a coordinated one is meant to pertain primarily to the high density COM traffic.

2. The use of a combination of essentially orthogonal and controlled frequency and time slotting gives the high multiple access performance efficiency of the waveform design. The PN portion does not significantly contribute to this but is included for multipath reasons primarily.

3. The use of both frequency and time slotting combine to give the system a high dynamic range. For severe dynamic range situations (the worst case near-far problem) the provision of some time slotting allows time separation of these interferers, whereas for simultaneous use of multiple frequencies, interference could occur because of equipment nonlinearities (assuming the tunable receiver front end cannot cope with the situation). On the other hand, time persistence interference can occur in the near-far situation for a time-slotted design, due to the propagation delay, the multipath environment, and equipment filtering plus nonlinearities. Choice of wide enough time slots alleviates these problems. However, for an all time-slotted design long frame times are required, even though not all users are exposed to this near-far interference at the same time. The combination of frequency slotting with time slotting, with coordinated use of the slots, should adequately satisfy the near-far problem for those users affected whereas users with less severe dynamic range requirements are not penalized by long waiting times.

4. Transmitter and receiver equipment timesharing is facilitated by the time slots in the waveform design, though to a certain extent in principle, one could use time compression techniques without going to TDMA slots.

5. Frequency slotting in combination with error correction coding improves anti-multipath performance for short delays. That is, the effect is made to be more of a fading one, which is easily handled by the error coding, assuming bit interleaving is also implemented.

6. Frequency slotting facilitates sync acquisition in accordance with system design, i.e., with digital matched filters, the clock rate and bit storage requirements are less, whereas with serial search methods the number of bits uncertainty is less.

7. Relative to impact on NAV measurement accuracy, suitable accuracy can be achieved with whatever amount of frequency slotting (one uses sidetone ranging) or PN (one ranges on clock phase) that is dictated by other system considerations (both PN and sidetone ranging are indicated here). On the other hand, pure time slotting depends on rather high peak powers that may not easily be achievable. A better NAV design in the latter case would be to combine PN spreading and/or frequency slot sidetone ranging with the time slotting.

8. The combination of a powerful error correction code (sequential decoding) with binary modulation (DPSK) gives a low E_b/N_0 for bit detection, which is desirable for the power-starved satellite downlink.

In the candidate FH/PN/TH waveform design here, the goal of a common waveform is met in the sense that a basic or elemental waveform is used, unchanged, for all CNI applications. However, the formatting and multiplexing for the different applications is not kept the same, the reasons being system optimization in each case. These variations do not have a significant impact on complexity - the black boxes required are still minimized.

Refer to Figure 4-11 for a diagram of the basic signal structure involved. It comprises a burst, which is 60 microseconds long (based on reducing the effect of system ringing in near-far situations due to multipath persistence) and contains 6 digits, each 10 microseconds long. These digits represent different information bits since they are selected from an interleaved pattern (so that digit errors for a bit in the channel will be random and independent) of error correction-coded bits. Since rate-1/2 convolution encoding/sequential decoding is used, 2 digits represent a bit (but the digits

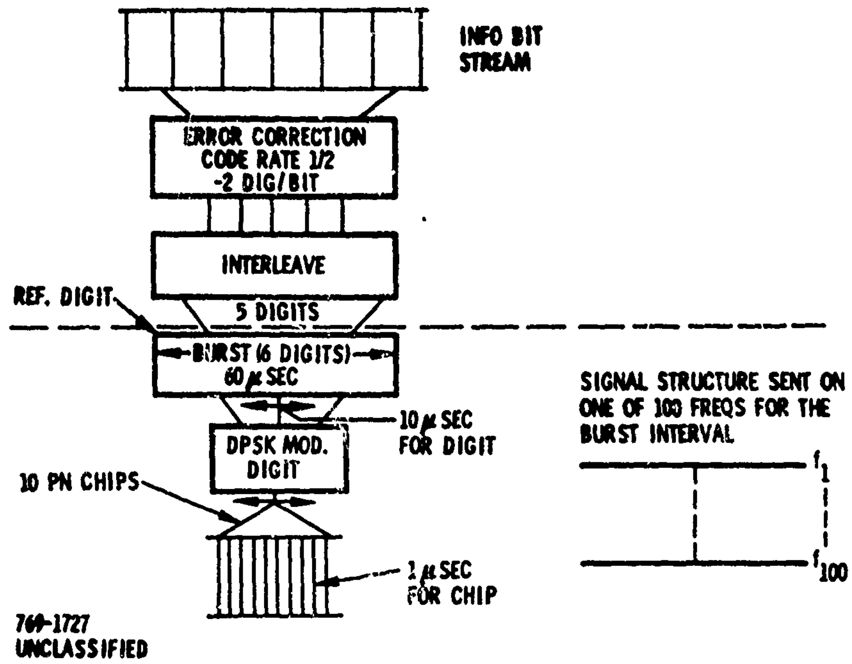

Figure 4-11. Basic FH/PN/TH Signal Structure for All Functions and Modes.

represent different bits in the transmitted burst). However, a phase-reference digit is included as one of the 6 digits in the burst, so that the latter actually contains a total of 2.5 bits.

On the other hand, to reduce the inefficiency involved in sending a reference digit in every burst interval, the coherent hopping capability discussed subsequently in Section 4.2.1.10 can be used so that a phase reference digit need be sent less frequently. In effect then, a burst of 6 digits will contain 3 bits of information rather than 2.5 bits. The higher value is used in the illustration cases for the succeeding figures. One can, however, scale the parameters accordingly if it is desired to send the reference in every burst.

The digits of the burst are DPSK modulated on a carrier. Each DPSK modulated digit is also quadriphase modulated by 10 PN chips (the PN is primarily to combat inter-symbol interference due to multipath). For 10 microsecond long digits we now have 1 microsecond for the duration of each PN chip. The bandwidth of the resultant signal is 1 MHz. The entire burst of digits is transmitted on the same frequency, which would be one of 100 possible frequencies in a 100 MHz bandwidth. This burst structure, as

indicated in Figure 4-11 is the basic signal structure for the coordinated FH/PN/TH waveform design to be used for all the functions and modes.

4.2.1.1 Direct COM

The formatting of this basic waveform (where formatting means the nature of the sequence of successive bursts in a message) can be made different for different functions and modes to optimize corresponding performance. Figure 4-12 indicates the signal format for direct mode 2400 bps COM. Since a coordinated waveform system design is involved (to achieve high capacity and efficiency of equipment time-sharing), the transmitted sequence for each access signal must be sufficiently long and/or allow enough guard time so that the propagation delay differences among the different access signals do not cause intolerable mutual interference. The signal format in Figure 4-12 indicates a guard time of 0.75 milliseconds and an on-time sequence length of 20 of the 60 microsecond bursts, giving, therefore, a total of about 2 milliseconds (sum of guard and on-times) for sending 60 bits. This entire sequence is sent on one frequency (out of 100 possible), with a change in frequency allowed for the next sequence interval, so that the hopping rate is 500 hps. To obtain a data rate of 2400 bps,

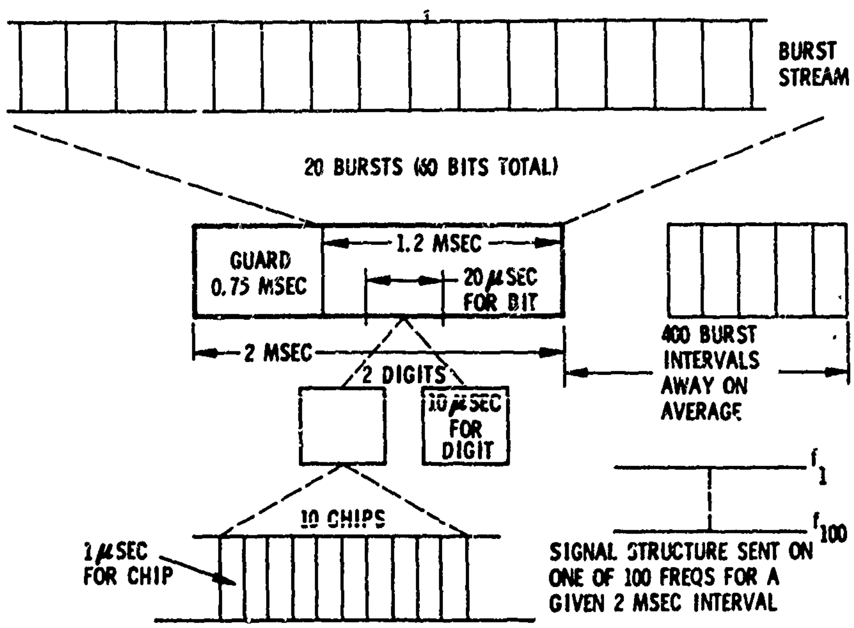

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Figure 4-12. Transmitted Signal Format for Direct Mode 2400 bps COM.

the frame time for the burst sequence transmission per signal access is 20 milliseconds. The corresponding time duty factor then is 20 to 1. However, because of the guard-time and interleaving of signalling slots, only 10 different 2400 bps signals are allowed on a frequency-time pattern on a TDMA basis. With 100 frequency-time patterns one then has up to a total of 1000 such signals. This scheme, however, presumes the ability to orthogonally separate out the different signal waveforms by use of coordination procedures and slot assignments based on NAV position. As an alternative to this tight control one would back-off in capacity from the 1000 total so that with a reduced occupancy of frequency time space, the statistical probability of mutual interference is at a tolerable level.

Figure 4-13 indicates the formatting for a possible system requirement of 75 bps signals in direct mode COM. Coordination of signal accesses is still assumed. Therefore the guard time and sequence length of bursts (20 of them) is the same as for 2400 bps COM in Figure 4-11. One also stays on a given frequency for the entire sequence length. However, the frame length for a given 75 bps signal access is now 32 times longer or 768 milliseconds (about 3/4 seconds). A corresponding greater number, or specifically 320, accesses are now allowed on a TDMA basis on a frequency-time pattern. They can coexist with the rest of the 2400 bps direct mode COM environment, since it uses the same signal structure parameters (as in Figure 4-11), but occupies one of the 100 total frequency time patterns.

4.2.1.2 Random Access IFF and Signalling

Figure 4-14 indicates the formatting for random access IFF and signalling functions. This function is defined here as the ability for different and multiple terminals to send short messages without prescribed slot assignments, and, therefore, without coordination of waveforms with each other. Its usage would be for those operational situations where random access type of IFF may be desired, or where supervisory signalling is required so as to get organized into coordinated nets (this would be for cases where one must allow for more addresses than there are slots, or where knowledge of slots is simply lost or does not exist). All these applications are lumped together in the random access function in Figure 4-14. It should be noted, however, that the suggested operational use of it for IFF does not preclude use of coordinated 75 bps surveillance type nets (as in Figure 4-13) for those IFF applications where it is both operationally feasible and desirable. Nevertheless, referring to Figure 4-14, it

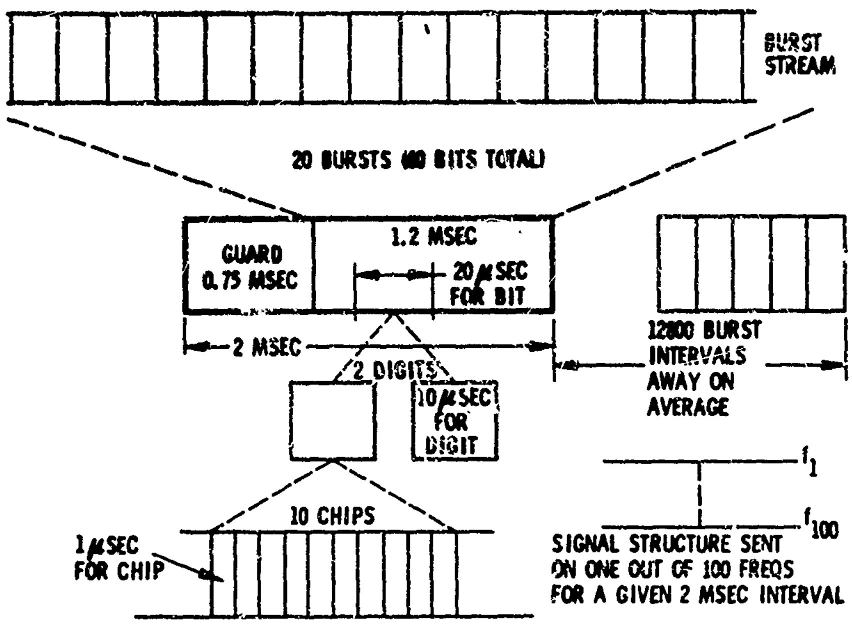

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Figure 4-13. Transmitted Signal Format for 75 bps Direct Mode Net

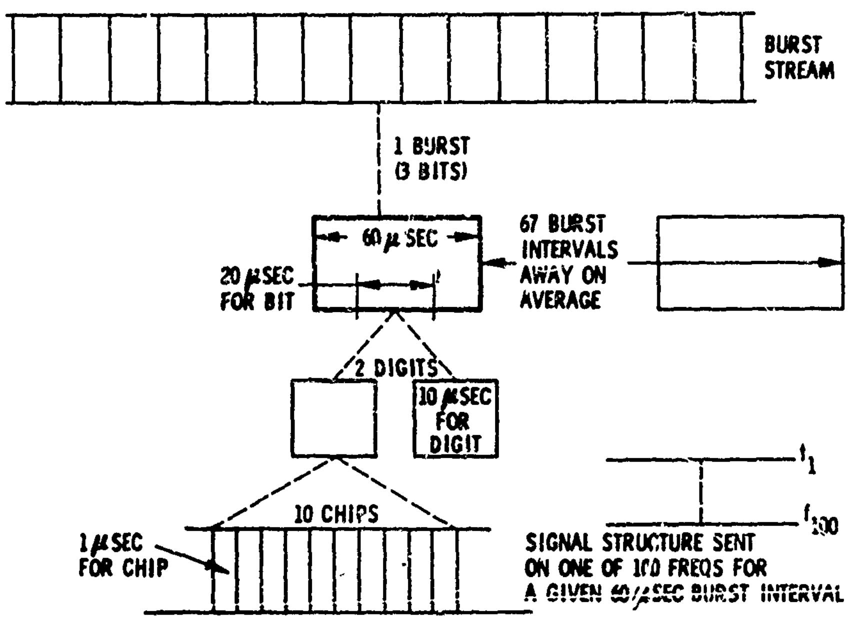

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Figure 4-14. Transmitted Signal Format for Random Access IFF and Signalling.

is seen that the same signal structure, but with the 60-microsecond bursts sent out individually (not in a sequence) is used. The frequency is changed from burst to burst (this randomizes the errors in the short IFF/signalling message so that error correction coding can be effective in achieving reliable detection). The spacing between successive bursts for a given signal is 2 milliseconds or greater, giving, therefore, the same frequency hop rate requirement as for direct mode COM, i.e., 500 hps. For a 48-bit signalling message, 16 of these bursts must be successively received. The waveform design here allows for repetition of the signalling message transmission and for guarding of two addresses by a terminal; these are, an individual address based on NAV position for the case of IFF or an identity address for the case of COM signalling, and a universal address which all terminals would detect.

By allowing for use of all 100 frequency-time patterns, but for a fraction of the time, a capability of up to 150 simultaneous random-access signals, as in Figure 4-14 can be accommodated. This assumes that the coordinated COM accesses are isolated from the random access IFF/signalling function here, so that interference between the two groups does not occur. As such, there is random-access mutual interference only among different accesses within the IFF/signalling function group. The design here would allocate a time slot or slots for this signalling function (16 percent of the time for this is illustrated in the design here). If an aircraft did not have timing information needed for this purpose, it is presumed capable of getting timing from the NAV information he has.

A possible alternative to this design would not orthogonally separate the coordinated COM accesses from the random access signalling function, but would simply overlay the two. A back-off in coordinated mode COM capacity from the maximum would be required to allow for the random access mutual interference. As for the added interference to the random access signals due to the coordinated COM nets, this is discriminated against by the fact that random access signalling would now have access to all the FT cells, not just to 16 percent of them. Though efficiency is degraded somewhat by this overlay technique, it has the advantage that knowledge of both timing and slotting is not required by the random access signals. In fact, the random access signalling function can be used to transfer timing as well as to obtain slot assignments.

4.2.1.3 Landing Mode

Figure 4-15 indicates the formatting for the ILS mode. Active transponder ranging and beam rider techniques are presumed here. For ranging, the transponder demodulates (range to the transponder must be known in a coarse sense for this purpose, i.e., to within a 1-microsecond PN chip width - this appears plausible based on NAVSAT position information), digitally delays for a prescribed time interval (as discussed in Section 4.2.1.10.2), remodulates, and retransmits on the same frequency on which is received. All aircraft in the landing zone are coordinated in waveform usage so that they would transmit only in every other slot, and allow the transponder to retransmit in the intervening slots. Slot assignments are made to the aircraft upon entering the landing zone. A 1-millisecond time interval comprising 16 bursts of the basic signal structure is used for the transmit and receive functions. This is wide enough to minimize the effect of mutual interference between slots. The total time for receive and successive retransmit at the transponder then is 2 milliseconds. The

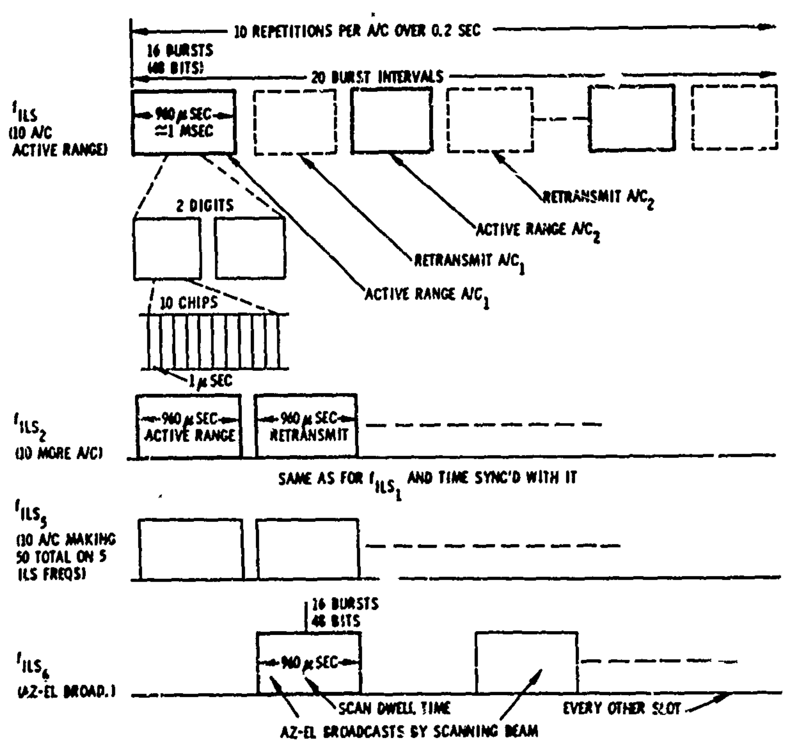

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Figure 4-15. Signal Format for ILS at Transponder for Active Ranging and Beam Rider

frequency is held constant over this interval and changed from interval to interval, e.g., 500 hps in the direct mode COM for aircraft outside the landing zone. The formatting design indicated in Figure 4-15 illustrates the case of 10 aircraft that are coordinated on a TDMA basis on each of five frequency time patterns so as to accommodate 50 aircraft in the landing zone with a fix rate of five per second for each aircraft, each at a different possible frequency in the 100-MHz bandwidth. Since coherent sidetone ranging discussed in Section 4.2.1.10 is used, a range accuracy of 6 feet is achievable in the presence of just Gaussian noise. Furthermore, the repetition scheme would mean up to tenth-order frequency diversity against interference such as multipath, provided its differential path delay is not too short (the coherence bandwidth must be less than the 100 MHz range signal bandwidth by the order of diversity considered - actually fourth order diversity, rather than tenth order diversity, should give adequate performance). Superposed on each ranging carrier transmission are 48 bits of information (16 bursts of the basic signal structure). For one repetition every 20 milliseconds this means 2400 bps. Therefore, a two-way voice link between the air traffic controller at the ILS transponder site and each aircraft in the landing zone is also provided by the ranging signal.

Now the ranging scheme in Figure 4-15 would be used whether or not the ILS transponder has an omni or directive antenna. However, given the efficacy of a 0.1 degree (elevation) by 0.2 degrees (azimuth) pencil beam at the transponder, and that traffic control knows the position of each aircraft (up to 50 in the landing zone) so that it can point the beam in the desired direction, it is desirable to have a pencil beam track each aircraft in the landing zone. There would actually be 5 such beams (one for each of the five frequency time patterns in Figure 4-15 that would be time-hopped from aircraft to aircraft of the 10 assigned to a beam, dwelling on an aircraft for the 2-millisecond basic rangi interval. This pencil beam would further discriminate against multipath (including t multipath for which we have no frequency diversity in a 100 MHz bandwidth) and other interference when a ranging signal is transmitted.

Assuming the less desirable alternative* that a tracking pencil beam concept is not used, the beam rider information, in terms of azimuth and elevation, is provided by broadcast step-scan beams that periodically cover all the possible angles for landing in the landing zone. When an aircraft intercepts one of these beams (threshold setting will

*Ground tracking may have undesirable system implications, in contrast to beam rider schemes not involving a tracking requirements.

discriminate against side lobe signals plus the fact that side-lobe-cancellation techniques will also be employed), the detection of the broadcast information will provide its azimuth or elevation coordinate. Figure 4-15 indicates this information as transmitted on a separate frequency-time pattern from the ranging signal which uses an omni antenna, and time synchronized so that the broadcasts occur in parallel only in the transponder range retransmit slots. The assumed dwell time of each beam is 1 millisecond. For this purpose multiple fan beams in both azimuth and elevation are required, the number depending on the angular resolution and volume of space required. It would be convenient to TDMA the azimuth and elevation scans so that each occupies 0.1 second for a total ILS fix interval of 0.2 seconds.

An alternate means of getting range information, as contrasted to the coordinated-waveform method described above, is to use the random-access IFF/signalling function mode. A $t_1 - t_2$ transponder concept would, of course, still be used (the transponder retransmits on the same frequency as the received signal but with a time delay). The ILS multiple-access capacity would not be as efficiently used now (in a bandwidth sense) as with the coordinated waveform design. However, it has the operational advantage of not requiring knowledge of a given slot assignment. Actually, since both a coordinated waveform capability and a noncoordinated waveform capability (as per Figure 4-14) are provided in this design, either method of ILS ranging can be used, depending on the operational situation. The coordinated waveform method is preferred.

4.2.1.4 Remote COM

Figure 4-16 indicates the formatting for the 2400-bps remote-mode COM, i.e., the satellite mode. Considering all COM that goes via the same satellite, the propagation delays for each signal should be known rather accurately (from the NAV data) so that insignificant guard times between signals are required. In this case it is desired to interleave the individual 60-microsecond bursts for each signal, rather than send sequences of bursts for each signal, so as to decrease potential repeat-jam vulnerability on the up link from any one site. Therefore, as shown in Figure 4-16, a single burst, of the same basic signal structure, is sent out for each signal with a frame time of 20 burst intervals on a given frequency-time pattern. Different signals are sent on 10 frequency-time patterns in parallel for a total illustrated capacity of 200. These signals are, of course, all coordinated. Satellite processing is used, as described later.

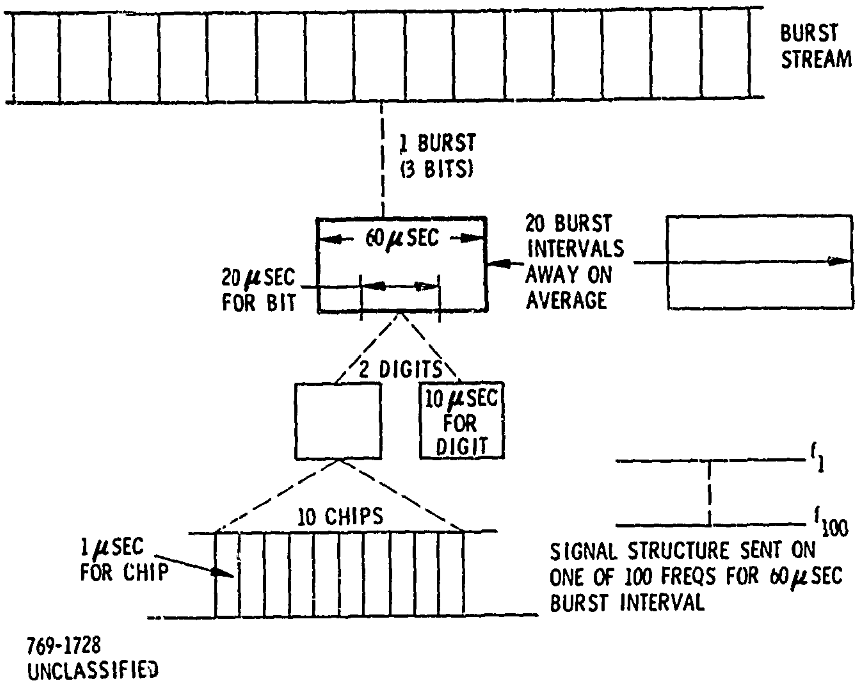

Figure 4-16. Transmitted Signal Format for Remote Mode 2400 bps COM.

4.2.1.5 NAV

Figure 4-17 indicates the formatting for the navigational satellite mode. In this case a sequence of 60-microsecond bursts of the basic signal structure are sent out consecutively from each NAVSAT. NAVSAT data is superimposed on these transmissions so that a 48-bit NAV message is sent every 16 burst intervals, or in about 1 millisecond. This message is repeated in subsequent transmissions in the 1-second total interval that is presumed to be allowed for a NAVSAT fix. The frequency-time pattern used for each NAVSAT is hopped at the same rate as the rest of the system, i.e., 500 hps. Each NAVSAT transmission is uncoordinated with the other NAVSAT's. The group of down link satellite frequencies for all the satellites is, however, separated on an FDMA basis with the direct mode signals (because the latter can use up to 100 percent of the available slots and can be stronger by 86 db or more) and frequencies are exchanged with the latter on a coordinated but much slower basis than the basic hop rate within each group of frequencies (because of the large variable propagation delays to the satellites coordination in time with the direct mode is possible only at a slow rate; however, because the rate of change of Doppler frequency to the different satellites is less than a channel bandwidth, coordinating them with the direct mode in frequency,

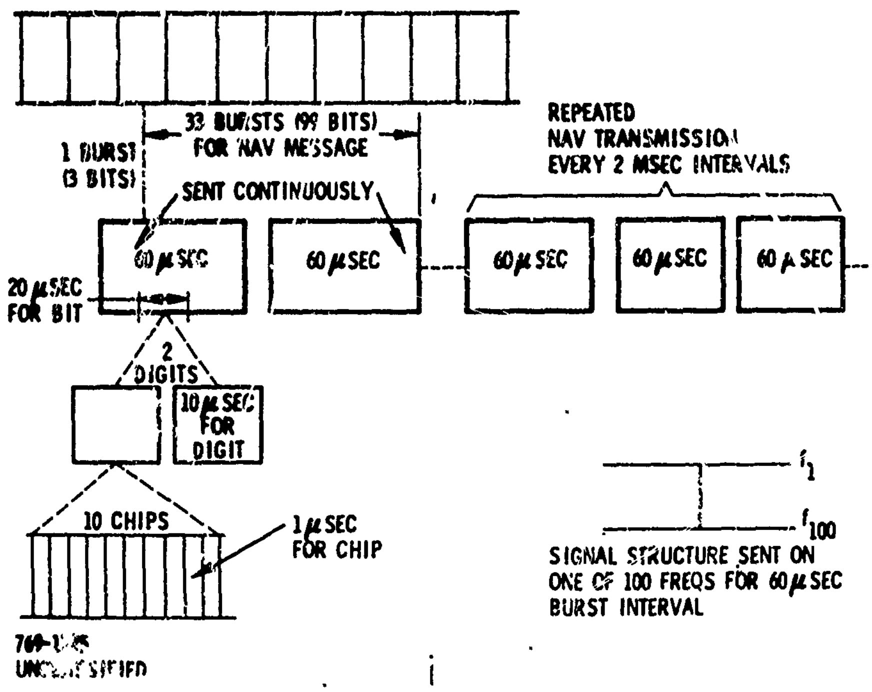

Figure 4-17. Transmitted Signal Format for Remote Mode Navigation.

i.e., on an FDMA basis, can be done easily). The NAVSAT signals are sent repetitively because:

1. Being noncoordinated with each other and with the satellite COM signals and also splatter from strong direct mode COM signals, the repetition processing gain is needed to discriminate against mutual interference.

2. Being coordinated with the direct mode mainly in frequency, yet desiring to timeshare the receiver processor with the direct mode (and also with the other NAVSAT's) multiple look possibilities at the NAV signal are needed.

This repetitive NAV transmission actually degrades efficiency in the order of 6 db or slightly greater since any one NAVSAT is processed only a fraction of the time (four NAVSAT's plus COM) whereas it transmits 100 percent of the time. Operational flexibility, however, is obtained in trade.

4.2.1.6 Multiplexing of Functions

Figure 4-18 indicates the multiplexing of the different functions (COM, NAV, random access IFF/signalling, and ILS) and modes (direct mode and satellite down link). Only the satellite up-link mode is not indicated, since it will be in a separate frequency band (a separate band is used because of the difficulty of operation in one common band). Figure 4-18 indicates a 100-MHz band for multiplexing of all the functions and modes indicated. A 1-second segment of time is also shown, to highlight the duty factors involved for the different functions (Figure 4-18 does not show the actual frequency-time patterns involved over this interval of time). In the frequency-time space the NAVSAT signals are coordinated with the other direct-mode functions on an FDMA basis. This means they are on separate frequencies. All four NAVSAT's are noncoordinated with each other in their use of the frequency-time pattern involving these frequencies. Their hop rate over them is the same, 500 hps, as for the rest of the system. However, the programmed exchange of frequencies used with the rest of direct-mode CNI is on a much slower basis because of the long variations in propagation delay to the different NAVSAT's. The NAVSAT's broadcast continuously on their

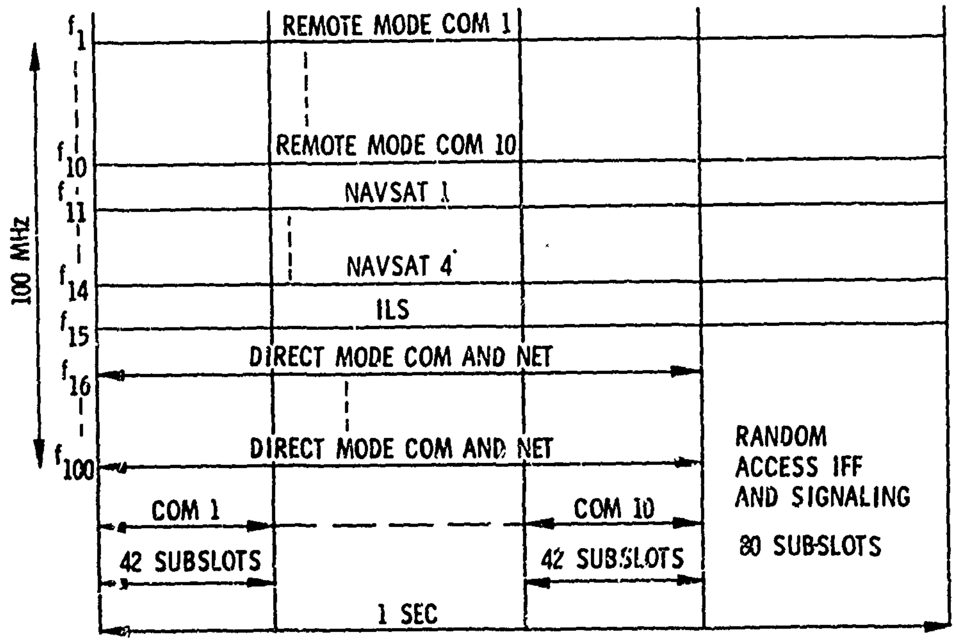

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Figure 4-18. Multiplexing of CNI Functions for Direct Mode and Remote Mode Down Link.

frequency-time patterns. The communication satellite down link is on a set of 10 frequencies which is also coordinated with the rest of direct mode CNI on an FDMA basis. The NAV and COM satellites can, however, share the same frequency. That is, though the different satellite signals may hop at the system rate in their group of frequencies, the group of 10 frequencies is exchanged with the rest of direct-mode CNI on a slower basis. The ILS mode now is on a set of 6 frequency-time patterns which is coordinated with the direct-mode COM frequency-time patterns. The random access IFF/signalling function is coordinated with the rest of the CNI system on a TDMA basis: that is, a specific time slot is provided in which all the frequency-time patterns are allocated to noncoordinated use for this function. The time slot is illustrated as 16 percent of all the sub-time slots. The direct mode COM function now occupies the rest of the frequency-time patterns and time slots exclusive of those used for the other functions and modes as described above. All patterns are intended to be occupied for maximum system COM capacity. On each pattern there are 10 different 2400-bps COM signals on a TDMA basis. They would be interleaved on a 2-millisecond interval basis rather than on a large group basis as shown in Figure 4-18. In parallel there are 80 such groups of allowed accesses for a total of 800 COM accesses. This is obtained by subtracting from the 1000 total system signal capacity the amount indicated for the other functions and modes. It should be noted, however, that each 2400-bps COM signal access can be replaced by an equivalent number (32) of, say, 75-bps accesses (or actually on a frequency-time pattern basis one would have 320), or a group of 2400 bps COM accesses can be replaced by an equivalent higher capacity access. For example, for 19,200 bps delta-modulated voice one could trade for 4 of the 2400 bps COM signals (this is based on eliminating the need for rate 1/2 error correction coding as discussed in Section 7.3.3 of Volume II but retaining the 10 db of PN coding).

4.2.1.7 Parameter Tradeoffs in Waveform Design

The above discussion indicates the multiplexing of all functions and modes for the case of maximizing system capacity (subject to the constraints of designing against intersymbol interference, needing to provide guard times, and so forth), providing a reasonable dynamic range performance, reasonable frame time, but limited somewhat in operational flexibility. Multiple tradeoffs are possible in the system waveform design with respect to these factors. The more significant ones are indicated as follows:

1. Operational Flexibility Versus Capacity:

Without changing anything else one can increase operational flexibility by reducing capacity by 5 to 10 to 1. For example, the NAVSAT signals and the down link communication satellite signals could be overlaid onto the direct mode COM signals on a noncoordinated basis, rather than coordinating them to be on separate frequencies as indicated above. One would also be able to overlay the random access IFF/ signalling function on a truly noncoordinated basis rather than having to have system timing so as to coordinate on a TDMA basis. In trade, so as to reduce mutual interference (the satellite signals would not harm the direct mode because of their lower power level, but the reverse would happen - similarly the random access IFF/signalling would not harm the direct mode COM because of its lower duty cycle, but the reverse would happen), total system occupancy of FT slots must be reduced from 100 percent to the order of 10 percent so that error-correction coding can be effective for data. Specifically, this would mean a reduction in direct mode COM signal accesses to something like 100.

2. Operational Flexibility Versus Intersymbol Interference:

By eliminating the 10 db of PN coding protection against multipath intersymbol interference and substituting 10 more frequency slots per signal, the improved operational flexibility indicated in 1 can be obtained without a reduction in capacity (1000 total for all signals). The idea is that by using redundancy in this way one achieves 10-percent band occupancy without a change in capacity - the mutual interference due to non-coordinated overlays thus can be corrected out. On the other hand, this scheme presumes strong intersymbol interference due to multipath will not occur (equal strength to a few db down is considered strong). The magnitude of this interference has not been determined so that any assumption is tenuous.

3. Frame Time Versus Dynamic Range:

If intersymbol interference can be ignored as in 2, and frequency slots be substituted for PN chips, one can also use the frequency-slot coding to reduce the frame time for a 100 time slot design. For example,

0.2 millisecond time slots could be used in the direct mode by constraining the choice of frequencies to 10 different ones in each of the 10 succeeding time slots that span a 2-millisecond interval (greater than the maximum propagation delay in the direct mode). Slot-overlap interference is thus reduced by such coding whereas the frame time for 100 time slots is held to 20 milliseconds.

4. Frame Time Versus Intersymbol Interference for More Time Slots:

By using 100 time slots in a frequency-time frame for 2400 bps direct COM signals, more terminals can be accommodated in a given "near-far" situation than with 10 time slots. However, the frame time (which concerns mainly system reaction time as well as buffer storage needs) is increased by 10 from over 20 milliseconds to over 200 milliseconds. Sync difficulties also increase if typically it takes more than one burst to obtain reliable sync. Also, the peak-to-average power trade requirement would now call for a 100 KW peak transmitter rather than a 10 KW one.

4.2.1.8 Civilian Derivative

Table 4-3 indicates the modification in parameters for a civilian version of the coordinated waveform design. Simplification, and, therefore, economy, is achieved for this design by eliminating complicated randomization and coding techniques (these include a coherent, random frequency hop synthesizer, PN generator and correlator, digital matched filter, sequential decoder, and need for a vocoder). Also factors such as having to generate large peak powers are eliminated. In trade performance is degraded, but not in all regards. That is, discrimination against non-Gaussian interference such as multipath and mutual interference is reduced, AJ is lost, dynamic range is more limited in special cases because there is no randomization of the interference, and the maximum civilian capacity is more limited. However, comparably, high NAV precision and ILS precision can be achieved in the presence of just Gaussian noise. It is noted, however, that given a civilian design without auxiliary aids such as an inertial platform, NAV cannot be obtained in range acceleration situations. Doppler, however, is not a problem because of the wide channel bandwidths used.

Table 4-3. Parameter Modifications for Civilian Use

- No Random FH/TH

    Fixed slot assignments for COM.

    Fixed pattern of 4 frequencies for NAV (different pattern for different NAVSAT's on FDMA basis) - sidetone range - simple frequency synthesizer.

    Fixed frequency-time pattern for random access IFF/signalling - patterns for different terminals assigned to uniformly fill FT space.

- For ILS, a 6-foot accuracy still obtained because 100 MHz used.
- Separate Civilian COMSAT used for COM but same NAVSAT's are possibly used. Signals are FDMA'd with each other and with direct mode.
- No PN coding of digits.
- Delta modulation (19.2 kbps) for voice instead of vocoder (2400 bps).
- No sequential decoding. For data, a simpler block coding is used.
- Peak-average power trade not implemented for burst waveform.
- With no inertial platform aid for NAV, range measurements are made only for no range accelerations.
- No digital matched filter for sync.

4.2.1.9 Functional Implementation

Figure 4-19 and 4-20 give transmitter and receiver block diagrams for the basic aircraft transceiver showing only aspects that are pertinent to the waveform concepts under discussion here. In particular, they tend to emphasize the equipment timesharing concept for the different functions and modes. In the transmitter it is seen that most of the major blocks are timeshared for the COM and IFF transmissions. Items timeshared include the coherent frequency hop synthesizer, RF power amplifier and antenna, IF amplifiers, PNG, DPSK modulator, and error correction coder. One can even timeshare the pseudorandom key generator if desired (and if security allows). A time-slot control program controls the timesharing of the equipments. It is further required that the COM and IFF data from the corresponding terminal devices be buffered and converted in format, of course. The timesharing of equipments indicated

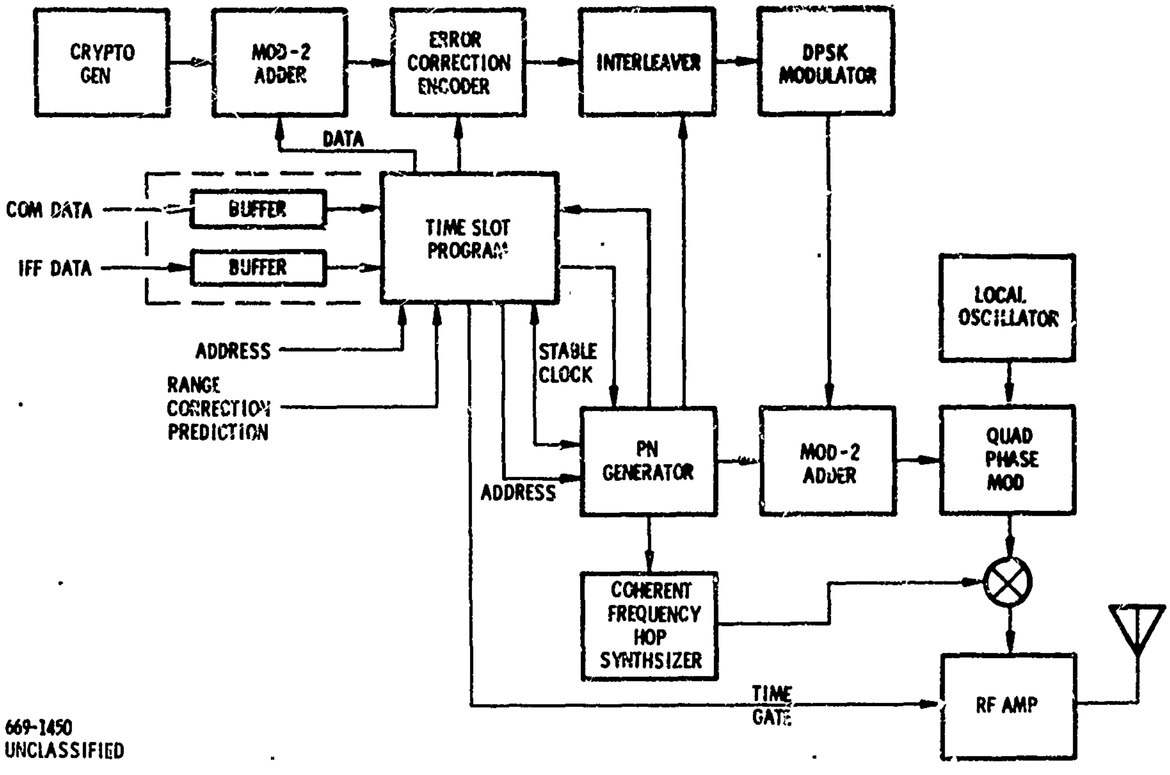

Figure 4-19. Transmitter (FH/PN/TH).

in Figure 4-19 also applies to the satellite mode as well as the direct mode. That is, it is intended that there be one of a kind of each equipment shown in Figure 4-19 whether one goes by satellite or direct (exceptions may be different antennas and possible RF tuning). Furthermore, it is feasible to transmit COM, say in the direct mode, while transmitting an IFF response in the satellite mode (simultaneity is, of course, defined over a given interval of time since we are timesharing equipments). The time slotting in the waveform design concept (1/10 for COM, and a separate slot for IFF) allows this. Coordination of direct mode timing with the satellite mode timing is not required at the transmitter for this purpose. The reason is the transmitter has multiple slot options to send IFF and can adjust this to avoid overlap with this COM transmission. In the receiver of Figure 4-20, it is seen that timesharing of one each of a multiple number of blocks (tunable RF front end, coherent frequency hop synthesizer, PNG, linear PN correlation, digital matched filter for sync, DPSK demodulation, and error decoding) is allowed for COM, NAV, and IFF because of the same aspects of time slotting in the waveform design concept. One could be receiving COM in the direct mode, and still get a NAV fix from the satellite mode (one of course shares the receiver for the

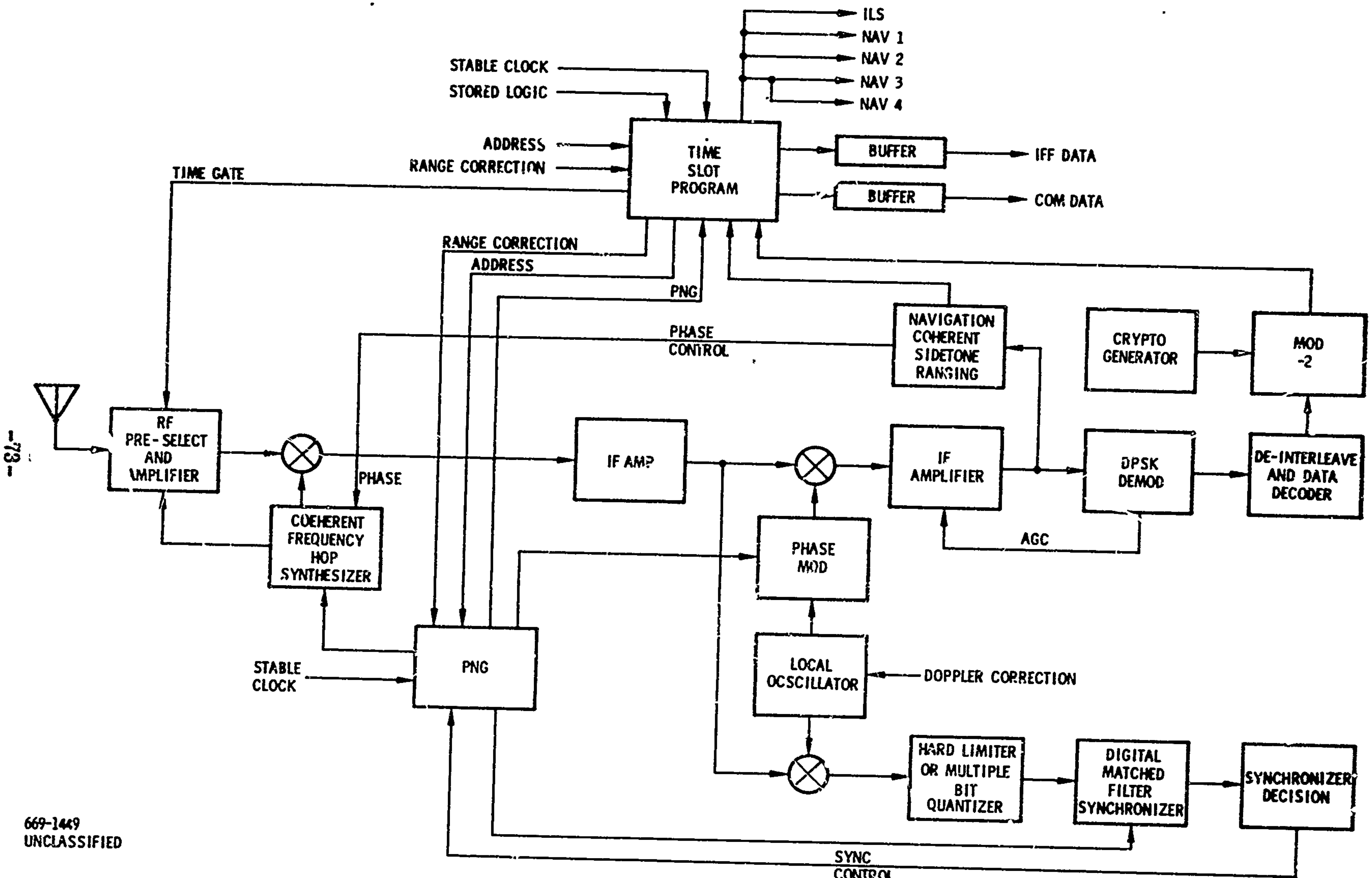

Figure 4-20. Receiver (FH/PN/TH).

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4 NAVSAT signals sequentially, and for the portion of time not overlapping the instantaneous COM slot, and the IFF function slot, as discussed previously). Similarly one can get IFF queries from either mode whereas COM is being received from the other mode. Coordinated timing of the direct mode with the satellite downlink signal is required for this purpose (by using the satellite as a central time reference for all modes, this becomes feasible). This is unlike the case at the transmitter where such coordination of timing is not required. Alternately just a duplication of processors is required if such timing coordination does not exist.

Also indicated in the receiver diagram of Figure 4-20 are processing functions not discussed thus far, but which are important parts of the system waveform concept here. Specifically, a digital matched filter is indicated as an aid to sync acquisition for COM, NAV, and IFF, while linear PN correlation detection is used after rapid sync with the d.m.f. The reason is that a d m.f. will have inherent implementation losses (under strong CW and other interference); e.g., 3 to 7 db or more depending on the complexity of implementation. This means any of reduced system processing gain or reduced capacity for COM, and longer IFF messages to allow longer detection time, if one also used the d.m.f. after sync acquisition. Also coherent sidetone ranging, for precise NAV, could not be used since RF phase information which it requires (after removal of PN modulation) is not available out of a digital matched filter. However, the d.m.f. performs well as an aid to sync acquisition for all the functions. The implementation loss is taken care of by allowing a corresponding increased sync processing time. Instead of seeking to acquire in one transmitted bit interval (this is 20 microseconds for a 50 kbps rate in a slot), one allows for acquisition ideally in 100 microseconds (= 5 x 20 microseconds). Actually to allow for a confirmation test and increased reliability, 200 microseconds or longer may be needed for sync acquisition. Considering that a slot contains 63 bits, this sync acquisition time is sufficiently fast, even for IFF, so as to acquire sync and also get information in one slot interval of 2 milliseconds. Under multipath fading conditions adequate power budget should exist in the direct mode to allow sync acquisition in the same time. (In the satellite mode for which there are power budget limitations, one supposedly is already in sync so that a digital matched filter actually would not be necessary.) For reacquisition, however, under limited power budget situations, and for special burst error situations, successive d.m.f. sync attempts could be made. In fact, multiple 2 millisecond intervals can be used. The implementation of the d.m.f., when used as a sync aid as in Figure 4-20 involves a

1 MHz clock rate and 100 PN chip bits of storage, all of which is within the present state of the art and is amenable to LSI implementation (see Section 5.2.4).

Another key processing function shown in Figure 4-20 is that of phase-coherent, frequency-hopped, sidetone ranging. This will be discussed further in subsequent paragraphs. Suffice to say here that it is the means of getting more precise ranging than that possible with 1 MHz PN, and yet be able to do it with the same channelized signal structure as for all the other CNI functions. Basically, the scheme involves frequency-hopping the tone modulation of the carrier (with PN modulation a 1 MHz channel bandwidth is occupied for each line of the modulated signal spectrum) over the full 100 MHz of system bandwidth. At the receiver after dehopping of the tone (and also PN decorrelation of the signal) the phase shift is measured to get range. The unknown phase shift that is measured will be due to just propagation delay since the instantaneous tone generated by the frequency hop synthesizer is made to have a deterministic phase (a phase-coherent, frequency hop synthesizer is used in both the transmitter and receiver of Figures 4-19 and 4-20, respectively). The phase shift measured due to propagation delay will yield precise range down to better than 6 foot accuracy because the tone modulation is hopped over 100 MHz.

Figure 4-21 shows the block diagram of the corresponding COMSAT that is suggested as part of this waveform concept. It indicates a channelized, processing satellite that involves 10 frequency channels (each 100 KHz corresponding to the waveform digit rate), frequency dehopping and rehopping. and PN decorrelation and recorrelation. Given 100 simultaneous COM accesses through the satellite, the coordination scheme is to be a uniform loading over the 10 time slots for COM, so that one has 10 accesses, each on a different frequency, in each of the time slots. The 10 IF filter frequency channels in the satellite also contain limiting so as to assure equal satellite power apportionment for each channel. In this way, that is with the combination of uniform time slot loading and power control in each frequency channel, efficient utilization of satellite power occurs. Since it is power starved this is, of course, an important consideration. Now the frequency dehopping and PN decorrelation in the satellite is to achieve maximum processing gain against an uplink jammer before he can grab satellite power. The dehopping of any given COM signal from 100 channels to one in the satellite gives 20 db discrimination, whereas PN decorrelation gives another 10 db. It is presumed that the aircraft transmitter can realize an almost 13 db peak-to-average power trade for being on for 1.25 millisecond in every 24 milliseconds

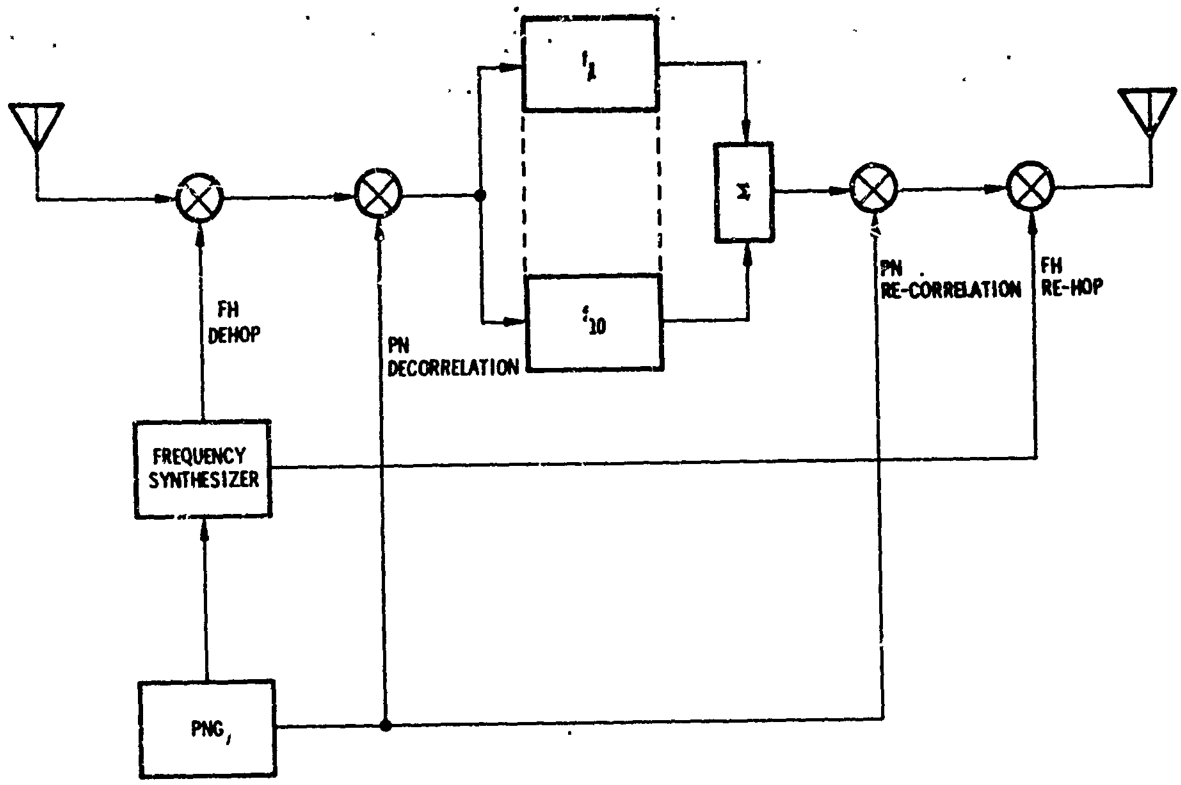

Figure 4-21. Satellite Processing (FH/PN/TH).

(on the average and considering IFF function interleaving) giving 13 db more of discrimination against an uplink jammer who cannot follow our time slot permutation strategy. The total waveform discrimination thus achieved against an uplink jammer who is in the same satellite antenna beam, and who would increase his effectiveness by grabbing satellite power in a power-starved system, is 43 db. It is noted that this is 3 db less than the ultimate received link processing gain because a rate-1/2 error correction coding is used, but decoding is not included in the satellite. The implementation in Figure 4-21 indicates just one frequency synthesizer to do both dehopping and rehopping (the retransmitted downlink signal must still be spread spectrum), and just one PNG to control the frequency synthesizer and do decorrelation plus recorrelation. This is for simplification in the satellite. Though this means that all aircraft transceivers must synchronize to these common devices, the most stringent requirement is only that relative to a 1 MHz clock PNG. A timing sync accuracy (1/20 of a PN chip) should be both adequate and feasible for transmissions to the satellite.

4.2.1.10 A Sidetone Ranging Scheme in Coordinated FH/PN/TH

In considering a design for precision ranging we note that the duty cycle of the signal used for precision navigation should be sufficiently high to meet the need for high update rates (two to ten per second) and the required extraction of velocity information as well as position information with high accuracy. This is not a great burden since the base points of a precision navigation system are generally associated with control centers which are the terminus of large numbers of communication feedlines. Thus, if we exploit the fact that range and range rate information can be extracted from the underlying FH/PN signal structure without concern for the data which is modulated onto it, we can enjoy the advantage of having simultaneous navigation and communication use of the same total signal (underlying structure plus data modulation) thereby gaining the desirable high duty factor for the navigation function without decreasing the number of time-frequency slots available for communication.

A solution to the precision navigation problem which appears extremely attractive is the use of coherent frequency hopping in the same coordinated FH/PN/TH structure described earlier. By requiring the phase of the hopping carrier to be deterministic (i.e., reproducible at the transmitter and receiver), the effective bandwidth of the signal is increased from the order of 1 MHz (which is the short term bandwidth stemming from the PN rate) to the order of 100 MHz (which is the total excursion of the hopping). Since the accuracy of range determination varies inversely with bandwidth, we have at once a two order of magnitude improvement. Furthermore, since we can coherently integrate over more than one slot time, some additional improvement is available beyond that available from smoothing independent measurements obtained from integration over individual time slots.

Before plunging ahead with the development and analysis of a fully coherent frequency hopping, we are well advised to consider the practical difficulty of achieving long coherence times at microwave frequencies due to the inherent noise in oscillators. Accordingly, it is best to assume that while coherent synthesis of the hopping offset frequency (over a 100 MHz) band may be substantially achievable, the microwave carrier on which it is conveyed will not be phase stable (remaining consistent within a radian) over more than a few milliseconds.

To aid in the explanation of this precision navigation concept we introduce the following numerical example. We suppose that a precision navigation transmitter,

which is also a communication transmitter, may be allocated an on-time of 1/8 of a channel. In particular we presume that it may transmit in two successive 2-millisecond time slots every 32 milliseconds. It is emphasized that these slots are not "consumed" by the precision navigation function since they may be simultaneously used for data transmission. Should the traffic density be greater for this transmitter, additional slots may be used; the regular pair of slots every 32 milliseconds is a minimum.

Consider that the transmitted signal is constructed by the combination of (1) a microwave carrier, (2) a frequency hopping subcarrier, (3) a pseudonoise sequence, and (4) data modulation. In this scheme the latter three are derived from a common clock. For example, the clock might be at 1 MHz which is multiplied by a factor of 1 to 100 to obtain the subcarrier. The clock directly triggers the PN sequence, which is divided by 10 to become the 100 KHz data clock and further divided to give word and frame timing. The synthesis of the subcarrier is done in such a way that regardless of the multiplication factor (supplied by the pseudorandom frequency hop programmer) the resulting sinusoid has a zero crossing coincident with the 1 MHz pulse.* At the receivers in the ordinary process of signal detection by extracted reference demodulators, the phase of the carrier is effectively measured. In this case the "carrier" frequency is the sum of the microwave carrier and frequency and the frequency hopped subcarrier frequency. By comparing the measured phase in the two successive slots, the difference in time between the 1 Mpps clock in the receiver and that at the transmitter delayed by the propagation time can be inferred with extreme accuracy.

The above description is now somewhat more formalized. Let

ω_c = carrier frequency

ω_1 = offset frequency during the first slot

ω_2 = offset frequency during the second slot

τ = time of propagation

ϕ = an unknown phase of the carrier

ϵ = error in receiver's estimate of τ

*A deviation from this condition is acceptable, provided it is systematic and is reproduced at the synthesizer in the user receivers.

Then the phase difference between the locally generated carrier and the incoming carrier during the first and second intervals are

$$[(\omega_c + \omega_1)(t - \tau) + \phi] - [(\omega_c + \omega_1)(t - \tau - \epsilon)]$$

and

$$[(\omega_c + \omega_2)(t - \tau) + \phi] - [(\omega_c + \omega_2)(t - \tau - \epsilon)]$$

The difference between these two phase measurements, subject to the assumption of phase stability of the carrier (the constancy of ϕ), is then

$$(\omega_1 - \omega_2)\epsilon$$

Since the offset frequencies are known from the hopping program, ϵ may be readily found and, with suitable filtering, fed back to adjust the phase of the 1 MHz clock. Since $(\omega_1 - \omega_2)$ can be a very large number, $2\pi \times 100 \times 10^6$, relatively coarse measurements of the differential phase can yield estimates of ϵ good to a fraction of a nanosecond.

One way of characterizing this scheme is to consider that it is a tone ranging system, with the difference that in usual tone ranging systems the tones are all available simultaneously, here they are available sequentially in a pseudorandom order.

While quasi-quantitative consideration of this scheme is encouraging, there are certain aspects of the signal extraction and processing which are difficult to analyze. The size of the frequency jump between the two slots affects the system sensitivity and appears as a coefficient in the loop gain of the receiver range tracking loop. Since this is a random variable, there are some questions of stability and loop dynamics. Because of this and other difficulties, it was decided to postulate a receiver mechanization and simulate its behavior on a digital computer rather than to attempt a closed form solution. Section 7.6.4 of Volume II presents the simulation results.

4.2.1.10.1 Mechanization

The transmitter for this coherent FH/PN/TH ranging system is block diagrammed in Figure 4-22. The master timing oscillator is to be a very stable source and synchronized to corresponding oscillators in the other emitters if a hyperbolic precision navigation system is used. This timing source, conveniently at 1 MHz, directly

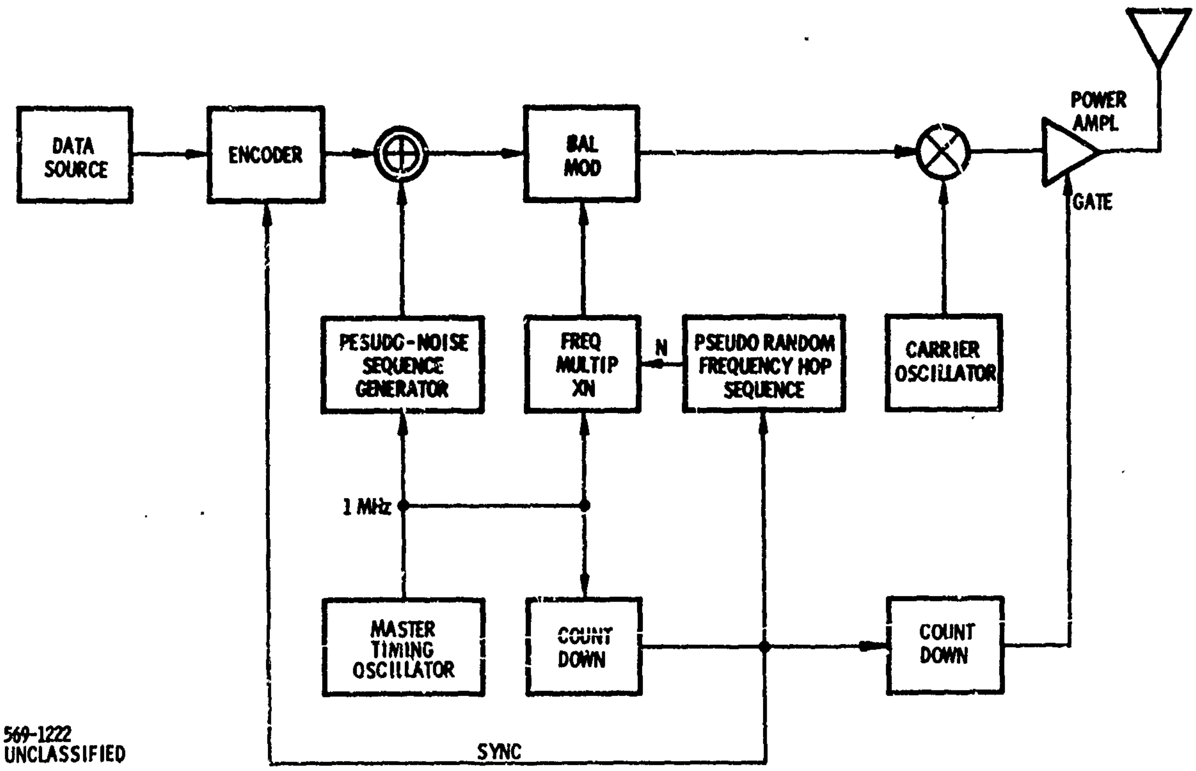

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Figure 4-22. Transmitter

runs the pseudonoise sequence generator which produces the stream for direct sequence phase keying of the hopped carrier. The same source, counted down, synchronizes the operations at the data input and error encoding. The counted down timing source also provides triggers for another pseudorandom sequence generator which produces the frequency hopping program. The synthesizer which produces the hopping subcarrier may be regarded as a variable frequency multiplier which multiplies the master 1 MHz by an integer (1 to 100) produced by the pseudorandom hopping programmer. The direct sequence is logically added Mod-2 to the encoded data and the result PSK modulates the subcarrier. This signal is then translated to the microwave and radiated.

A similar conceptual diagram for the receiver is given in Figure 4-23. After translation down from microwave, the hopping subcarrier is removed by a locally generated replica. This is formed in the same way as at the transmitter, except that the master timing oscillator is voltage controlled. After de-hopping, the direct sequence PN is removed by multiplication. The residue of the signal is now the data modulation on an IF carrier whose phase is dependent on any errors in timing. A

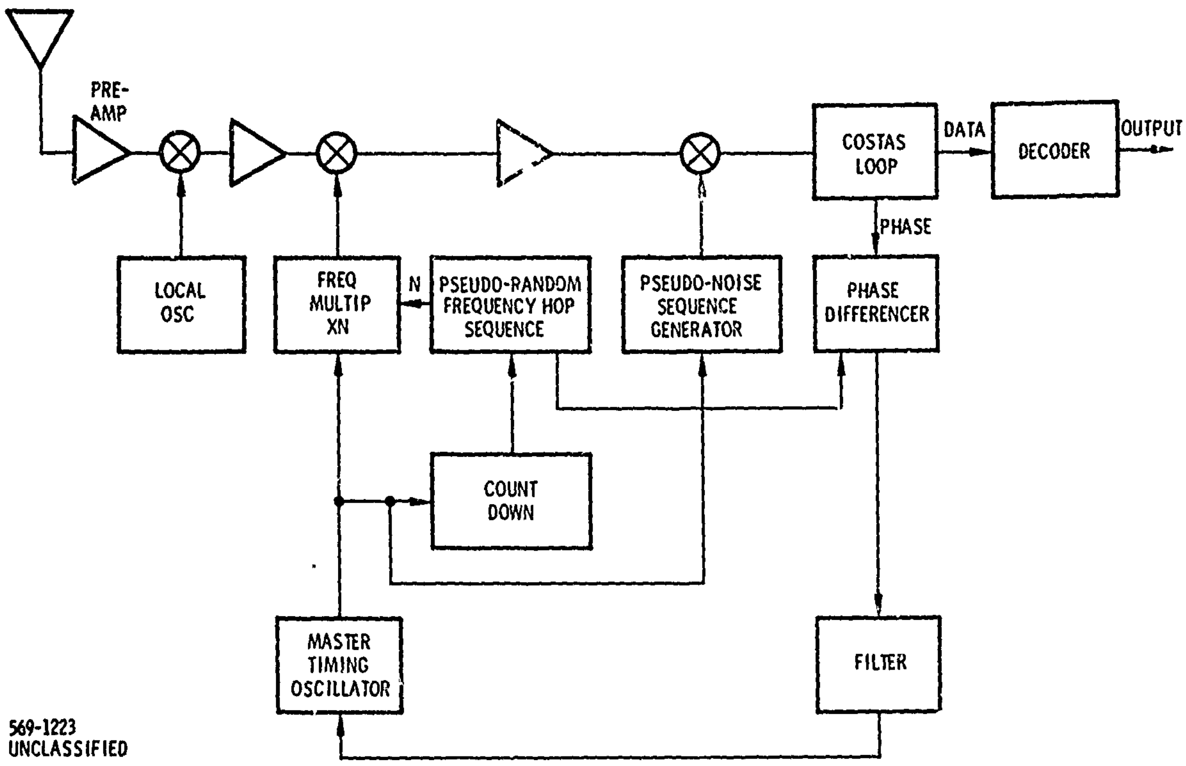

Figure 4-23. Receiver

Costas loop or equivalent can detect the information and incidentally provides a filtered carrier. The phase difference of this filtered carrier in the successive slots can be measured and, making suitable adjustment for the size and sign of the frequency step, an error signal is developed for correction of the master timing oscillator.

Synchronization, not indicated explicitly in the block diagram, can proceed as follows. The master timing oscillator of the receiver is time and frequency set at regular intervals, typically for an aircraft just prior to take-off. When trying to acquire a signal, this oscillator is slewed back and forth about its regular rate so that the frequency hop and direct sequence slew back and forth in time. When the timing comes within the width of the correlation peak of the PN direct sequence, an increase in energy will be observed at the Costas loop, or an auxiliary detector at that point. (Alternately, a digital matched filter enables faster synchronization.) The correlation peak for a 1 Mbps sequence is a triangle, 2 microseconds wide at the base. Upon detecting sync, the slew is discontinued and replaced by a small dither on the master oscillator which varies the timing slightly so as to produce a steering signal to center the timing on the correlation peak. After a short smoothing period, this centering can

readily be made good to within about 0.1 microsecond. The control loop may then be closed to utilize the error signal from the hopping in place of the energy variation due to the dither and the timing loop will then pull in to a fraction of a nanosecond. Section 7.6.4.4 of Volume II discusses the loop behavior in somewhat more detail.

An alternate implementation of the coherent tracking loop potentially more suited to time sharing is now described. Essentially, the coherent hopper is operated from a fixed frequency reference and the requisite phase shift $\Delta\omega\epsilon$ is introduced at another convenient point in the receiver. This implies that the time error is carried as an analog or digital variable which is available from the processing circuitry separately for each signal to be tracked. This time error is to be multiplied by the selected frequency offset $\Delta\omega$. Figure 4-24 shows the essentials of one possible receiver processing approach wherein the PN clock is derived by phase shifting the reference by the indicated time error ϵ, and the hopper output is phase shifted by $\Delta\omega\epsilon$ via a digital mechanization. Figure 4-24 may be compared with Figure 4-23.

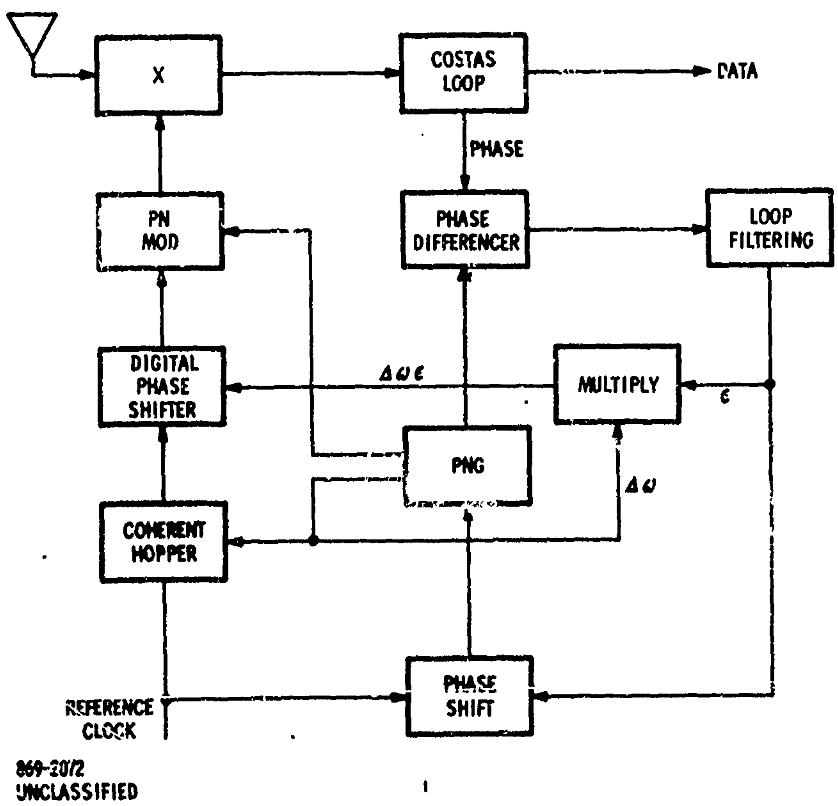

Figure 4-24. Digital Phase Tracking.

The quantization required for the digital phase shifting of the hopper output is now examined. A particularly simple implementation concept would utilize the quadriphase PN modulator already provided to quantize into 90 degree sectors by a logic transformation. Thus, the additional complexity for phase shifting is negligible.

The maximum error due to the quantizing (properly done) is $\pm$ 45 degrees, and the quantizing may be presumed to introduce phase jitter which can be randomized over successive hop intervals. Thus, if the jitter is uniformly distributed over $\pm$ 45 degrees, the rms value is 26 degrees, corresponding to $\sqrt{2} \times 26^{\circ} = 37^{\circ}$ for the rms value of the phase difference between pairs of successive hops. If the average frequency hop is taken to be roughly 50 MHz for a 100 MHz total bandwidth, this rms jitter corresponds to approximately 2 feet. It may be asserted that averaging over several successive hops will reduce the error of quantizing to a negligible value.

The logic transformation for digital phase shifting may be understood by noting that two PN chips are needed for quadriphase modulation to select one of four possible phases, and depending on the quantized value of $\Delta\omega\epsilon$, an appropriate rotation of the coordinate system is to be made.

4.2.1.10.2 Active Ranging for Landing Mode

A concept potentially useful particularly for the landing mode involves active ranging from the user aircraft to a transponder at the ground site. This concept differs significantly from range-difference schemes which lead to hyperbolic intersecting contours and consequent accuracy reduction due to geometry. With active ranging, a measurement of range is obtained directly, and accuracy reduction is eliminated.

If the transponder operates as an instantaneous $f_1 - f_2$ wideband repeater, in similarity with the satellite communication repeater, the range measurement scheme is evident, and the accuracy is determined by the aircraft equipment characteristics. However, $f_1 - f_2$ operation is not convenient for the direct mode, which includes landing as a compatible mode. For waveform concepts using time slotting, a $t_1 - t_2$ operation may be envisioned as an alternate. Now the transponder necessarily incorporates memory in its implementation.

The FH/PN/TH coordinated waveform utilizes coherent frequency hopping for a more accurate time of arrival measurement. The $t_1 - t_2$ transponding operation should be compatible with this. The purpose of this discussion is to show the conceptual

feasibility of such transponder implementation, at least for the case where any one signal interrogation is to be handled at any one time in a frequency slot. For the discussion, the transponder will be presumed to respond independently in the respective frequency slots provided, and these slots are hopped in accordance with the assigned code pattern.

In time slot t_1 the interrogating signal is received on the assigned frequency. A phase lock loop (or Costas loop) tracks the carrier phase as received in the transponder and arrives at a best estimate at the end of the time slot. The phase lock loop is presumed to retain this phase estimate until time slot t_2. (In an analog phase lock loop with high gain this can be implemented by holding the drive to the loop filter at zero after slot t_1 has ended and the received signal has disappeared.) It is assumed that retransmission by the transponder occurs in time slot t_2 on the same frequency. Provided that the frequency slot spacing is exactly the reciprocal of the time slot duration, the delay of an integral number of time slots simply introduces a multiple of 2π phase shift on the carrier. Hence, there is no inherent phase error introduced by this transponding process to prohibit application of coherent frequency hopping to achieve accurate range measurement.

In the above discussion, carrier Doppler has been tacitly omitted, and the received frequency is presumed exactly correct at the transponder. This will require the aircraft to implement an offset of its interrogating signal to arrive at the transponder with zero Doppler. The requisite accuracy may be estimated from the time delay involved. For example, with a delay of one millisecond, described in Section 4.2.1.3, an uncompensated Doppler offset of 500 Hz will introduce a phase error of π, and this is the bound for an unambiguous phase measurement. Thus, a Doppler error of roughly 100 Hz is suggested as the maximum with one millisecond delay; this corresponds to an allowed uncompensated velocity error of about 50 feet/second at 2 GHz. In addition, with this velocity error, random frequency hopping over the 100 MHz signal bandwidth introduces a random phase jitter on the measurement with a peak of 2 degrees for a delay of one millisecond in the transponder. This jitter is acceptable since it can cause a maximum range measurement error of 0.1 foot, at most.

4.2.2 TH/FH/PN HYBRID-COORDINATED WAVEFORM

The direct mode of operation for CNI has several characteristics that make a noncoordinated waveform an interesting alternative candidate to the coordinated

structure despite the necessarily reduced total multiple access capability. The direct mode of operation - consisting of communication and IFF - may be characterized by a large number of subscribers, operating on low duty. These subscribers will want immediate access and will have short transmission times. This mode of operation presents problems relative to coordinated operation. First, the coordination itself is cumbersome. Operational doctrine may be difficult to establish and maintain. Further, fully dedicated addresses used only intermittently make bandwidth usage very inefficient. For instance, if a 100 MHz bandwidth system employs an efficient coordinated TDMA whose users transmit at a 4.8×10^3 bit per second symbol rate, 20,000 users can be supported. If, however, these are dedicated addresses whose users operate one percent of the time, the average occupancy for the 100 MHz bandwidth is 200 users. The freedom from coordination constraints and the equalizing effects of possible low usage, then, lead to further consideration of noncoordinated waveofrms particularly for the direct mode. Definition of a TH/FH/PN hybrid-coordinated waveform is presented below, which uses coordination in the remote mode only.

4.2.2.1 Basic Waveform

The basic waveform to be employed in the uncoordinated system uses a 60 microsecond pulse duration and a duty factor of 4.8×10^{-3} resulting in a 12.5 millisecond frame time. The 60 microsecond pulse contains 60 signalling digits (for the 2.4 Kbit per second data rate with rate one-half coding). The pulse is further modulated by a 10 Mbps pseudonoise sequence making the total bandwidth occupied by the waveform 10 MHz. There are, thus, ten separate frequency channels in the 100 MHz bandwidth.

Frame synchronization exists to the extent that all users employ a frame synchronized to the satellite. Transmissions from each user, then are timed so that they arrive at the satellite at a known time relative to satellite time. Aside from frame timing, however, the direct users are entirely uncoordinated.

Each address determines a burst time and a frequency independently during each frame time. There will be times when the transmissions of two or more users

are superimposed during any single frame time. Since the occupancy of each frame is independent of the others, the interference will be independent from pulse to pulse; interleaving will be employed to combat the burst like interference.

The rationale for this waveform is as follows. To relieve the restriction of requiring complete coordination in the high activity direct mode, a low duty factor waveform is chosen to provide a significant co-channel occupancy capability. The pulse duration is made somewhat long to reduce the effects of channel dispersion causing pulse stretch and multiple pulses. The duty factor then results from a consideration of synchronization, data storage, and multiple access. The first two factors are enhanced by higher duty factors, the latter by lower ones. The final design then becomes a compromise between these factors.

The pseudonoise waveform is superimposed on the waveform for two reasons. First, it provides protection against multipath caused by intersymbol interference. Secondly, it provides a waveform of sufficient bandwidth to provide the navigation capability in a straightforward manner. In addition, the frequency diversity is used to provide resistance to signal fading and to make use of the full bandwidth available. The channel will therefore be naturally extendable in 10 MHz intervals - which need not be contiguous.

The multiple access capacity in the direct mode is computed in Section 7.3.4.1 of Volume II to be approximately 200 simultaneous COM and IFF signals under the worst-case assumption of strong signals but with the time-frequency orthogonality of 2000 slots being maintained. Up to 20 nearby (very strong) transmitters can be active considering the 200 time slots only and ignoring any benefits of receiver frequency selectivity. This multiple access capacity is based on allowable 0.05 probability of error per digit.

The antijam performance against broadband jamming was computed to be J/S = 39 db in the absence of simultaneous signals. There is a tradeoff between multiple access and antijam. Also, when a partial band jamming strategy is postulated, the maximum tolerable J/S is reduced several decibels. The precise amount depends on the data modulation scheme; for DPSK the J/S is reduced to approximately 37 db with a jam strategy hitting about 0.3 of the slots. See Section 7.3.1 of Volume II for further discussion on this point.

In the remote mode, the multiple access capacity with a noncoordinated waveform is significantly degraded in a power-starved satellite channel, as discussed in Section 7.3.5 of Volume II. For this reason, the waveform will be presumed coordinated for the remote mode. This leads to a relatively simple satellite repeater configuration, as will be seen.

The CNI system must serve a multiplicity of functions, each of which has at least slightly different criteria for performance. The basic waveform - a 60 microsecond pulse band spread over 10 MHz and hopped among 10 channels at a duty factor of 4.8×10^{-3} - is used, unmodified, for all functions. In the following, the application of this waveform to each of the functions is described, and operational block diagrams presented showing the basic functional implementation of the system. The description shows the high degree of commonality that is obtained in the implementation for the various functions.

4.2.2.2 Direct Mode

It is assumed that in the direct mode an aircraft would have to guard two voice channel nets and participate in a data channel net. The data channel will be considered continuous - so that synchronization is not a problem - and the voice channels are intermittent. Only a single channel transmission is required; however, the transmission must not interfere with the reception of data and navigation information.

Figure 4-25 shows the aircraft transmitter block diagram. Two data information sources are shown, a voice source and a data source. The data source represents the input from both the data link and the IFF responses. In the diagram, the data source is shown as being encoded whereas the voice source is transmitted without coding. The data source is encoded because the required 10^{-5} bit error rate can be obtained significantly more efficiently using coding. The voice information, if it is assumed that 2400 bit/second vocoded voice is to be transmitted, can tolerate a 2% or higher error rate without significant reduction in quality.* Therefore, no encoding is required in the voice channel. For commonality, the voice channel may be presumed to use 4.8 kilobits/second data rate without coding.

*R.W. Steele and L.E. Cassel, "Effect of Transmission Errors on the Intelligibility of Vocoded Speech," IEEE Trans. on Comm. Systems, March 1963, pp. 118-123.

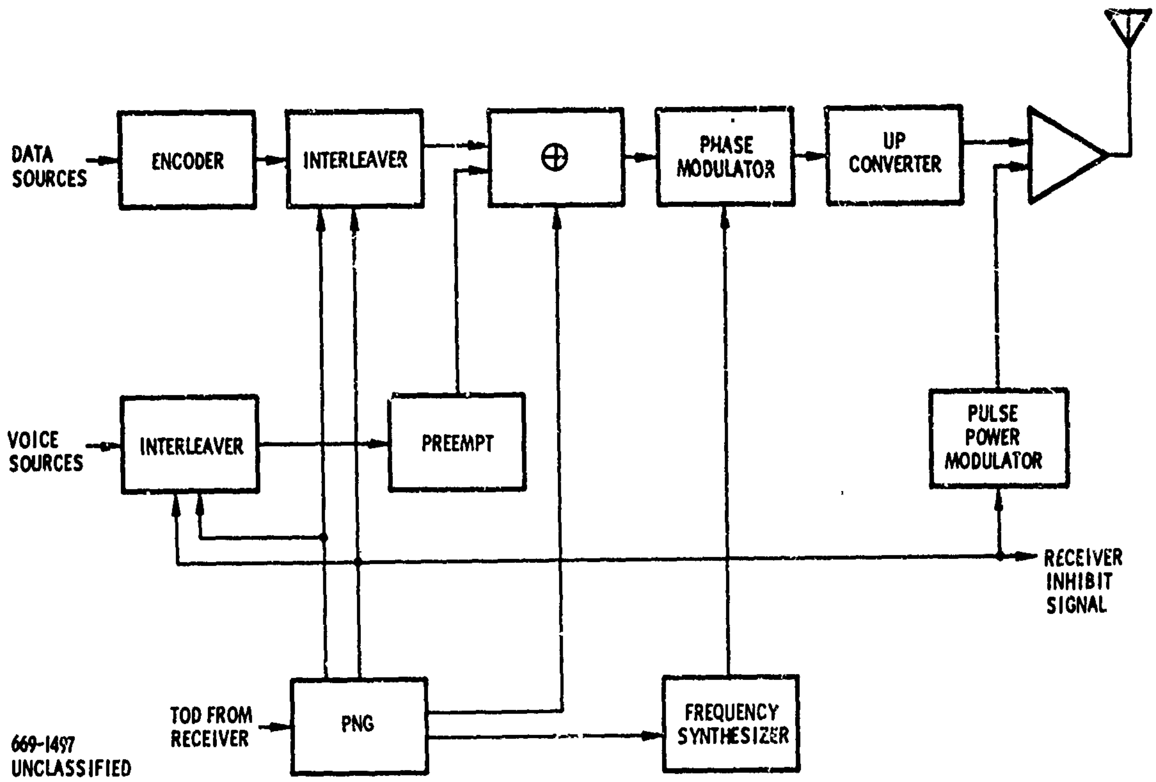

Figure 4-25. Transmitter.

Both information sources are shown in Figure 4-25 as employing interleaving in the channel prior to mixing with the modulation. This interleaver provides two services in both channels. The interleaver randomly distributes the bits over the extended time interval and provides the rate change required to change the 2.4 kilobit/second (4.8 kilobit/second in case of the encoded information) to the required burst rate for subsequent transmission. Two interleavers are required here, for, while more than one burst signal is never actually transmitted at a time, both channels may be required to transmit data simultaneously.*

The low duty factor of the waveform is the feature that allows this simultaneous operation in real time with sequential operation in the burst mode. In instances when there is pulse overlapping the data source is given pre-emptive capability over the voice source since the voice is more tolerant of errors.

* The interleaver in the voice channel may ultimately provide only the continuous to burst buffering. Because voice has a high tolerance for burst interference, the randomizing may not be required.

The outputs of the interleavers in Figure 4-25 are modulo-2 summed with the output of the PN modulator to provide a 10 megabit per second baseband signal. This signal is then used to phase modulate the output of the frequency synthesizer and this modulated IF is subsequently upconverted and power amplified in the pulse power transmitter.

As shown in Figure 4-25, a single pseudonoise generator provides the interleaver pattern, the pulse patterns, the pseudonoise modulation, and the frequency synthesizer command signal. The pseudonoise generator receives time of day clock from the receiver and generates all the above information by different manipulations of command words generated by the pseudonoise generator. An additional use of the pulse pattern generator signal is to provide a receiver inhibit signal to the switch in the front end of the RF circuitry of the receiver, isolating the receiver front end when the transmitter is on. The output of the transmitter then is a 10 megahertz wide PN modulated data stream having a pulse duration of 60 microseconds randomly placed within a 1.25 millisecond frame time. During instances of parallel transmission of voice and data, the duty factor of the transmitter will go up by a factor of 2 for the duration of the data message. An alternative to this variation in duty factor is to blank the voice transmission during the entire data interval.

It should be noted that in this transmitter only the interleavers are duplicated. The remainder of the circuitry is either used sequentially as in the case of the frequency synthesizer and all of the modulation and upconversion mechanization, or a single information source is operated upon in different manners to provide the different channel patterns as in the case of the pseudonoise generator. Operating in this time sharing manner makes the transmitter very little more complicated than a single channel transmitter.

The receiver circuitry required for the direct mode is shown in Figure 4-26. The RF section of the receiver implementation is standard in that there is an isolator switch, a broadband RF preamplification stage, and a down-converting stage that, because the output of the frequency synthesizer is employed as the local oscillator, reduces the original 100 MHz bandwidth to 10 MHz prior to going into the pseudonoise correlator. In the pseudonoise correlator, the output of the local pseudonoise generator mixed with the local oscillator is correlated with the received signal providing a narrowband IF that is subsequently synchronously demodulated. The output of this

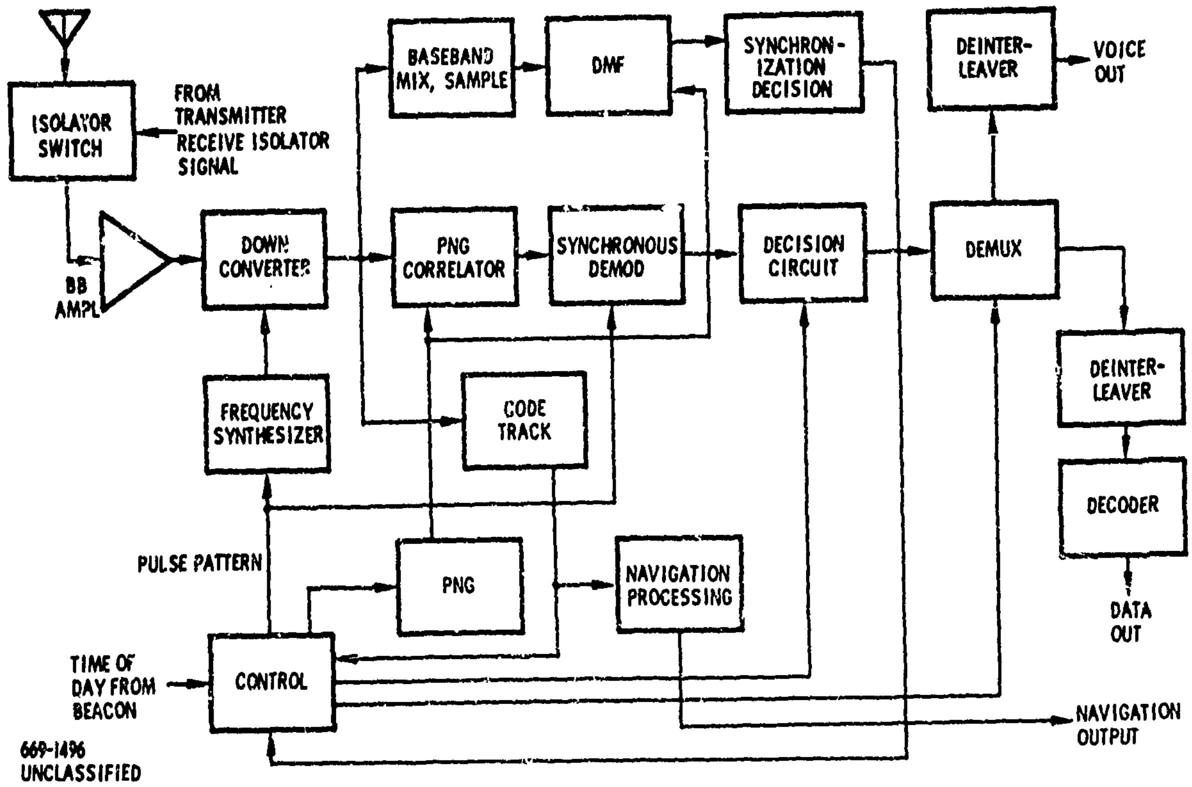

Figure 4-26. Direct Mode Receiver.

demodulator is used to make all of the data decisions for the system. In all three stages - down conversion, pseudonoise correlator and synchronous demodulator - the implementation is employed sequentially to provide for the different functions that are to be received.

In the direct mode, the system is required to search for two possible voice addresses and track the data address. When a signal has been acquired, the search mode for that channel is abandoned and the receiver is required to monitor that channel only 1/200 of the time. The previous decision to preempt the voice in favor of the data signal is also employed in the receiver, however.

It is during synchronization, when the receiver is searching for the two voice channels, that the time requirements on the time sharing equipment are most severe. The data channel is being tracked continuously so it requires only one 60 microsecond pulse duration out of the frame time. The voice channels, however, suffer from an uncertainty in time of arrival of 1.8 milliseconds. The synchronization procedure for

these channels then is to switch to the required frequency and pseudonoise code phase at the earliest possible time of arrival of the desired pulse (the known frame time allows this). The output of the down converter is switched to a detector shown in Figure 4-26 that provides the baseband output of the in-phase and quadrature products of the received signal, sampled at the system sampling time. These inputs are fed to the digital matched filter along with the pseudonoise code vector state. The output of the digital matched filter, which constitutes the sum of the squares of the inphase and quadrature matched filter channels, is processed by the sync decision circuitry.

The synchronization circuitry abandons the long search aperture on the results of a single pulse. Subsequent frames are examined only in a position corresponding to the detected pulse arrival. This is continued throughout the message or until subsequent processing dictates the original acceptance invalid. When operating in the search mode the decision circuitry must be dedicated to a single channel for the maximum time delay plus the pulse duration. In this case the dedication would last 1.85 milliseconds. Because the frame time is 12.5 milliseconds, a requirement for synchronization of two or more signals simultaneously can not be handled well by a single synchronization circuit, which must be dedicated to each signal for 15 percent of the time. Two synchronization circuits operating independently in parallel can be provided. Then, for example, with three signals, the probability is .0225 that both synchronization circuits have been preempted from serving the third signal.

When a synchronization decision has been made, the controller stores the time of arrival of the desired pulse and subsequent frames are switched through the synchronous demodulator into the normal data circuitry. As noted earlier, the decision circuitry is used sequentially for the voice and the data mode. A demultiplexer, therefore, delivers the information to the respective deinterleavers and the output of the deinterleavers is continuous data, either to the voice output or to the decoder for the data.

4.2.2.3 Navigation Mode

The navigation signal is completely asynchronous to the data signal and employs exactly the same format. Burst times, frame rates, and bandwidth are all identical. This implies a hypothetical navigation satellite RF source capable of high peak power transmission in accordance with the low duty factor of the signal. The information rate in each channel is low, in the vicinity of 100 bits/second; each of the four navigation

satellites must be continually tracked by the receiver. The receiver diagram shown in Figure 4-26 provides all the functions required to track the navigation signals. In block diagram form, the navigation receiver function is identical to the direct communications receiver function. The synchronization decision circuitry is seldom used since the navigation signals are tracked continuously; therefore, the navigation function can be performed with the remaining communications circuitry, which is used on the 1/200 duty factor basis. However, here there are four signals to be tracked continually. The navigation satellite will provide update information on the average of once every second. Thus, averaging over 100 frame times, a significant amount of overlap can be tolerated by the navigation signal.

4.2.2.4 IFF Mode

The IFF receiver must be able to discriminate at least three receiver queries and respond within a 1 second time interval. It is presumed that all information directed toward a specific aircraft will have a single unique address. Therefore, the IFF receiver need guard only one channel for acquisition. Synchronization detection is similar to that of the direct mode communication, and parallel circuitry can be shared. The difference here is that a multiplicity of detected signals must be provided for and that since this multiplicity will be constrained to a 2 millisecond interval, it is very likely that parallel processing for the three individual accesses will be required subsequent to synchronization. Again, a high redundancy short constraint length code will be used with interleaving to provide message integrity in the IFF mode.

4.2.2.5 Remote Mode

The remote mode will be operated in a coordinated manner. A separate (from the direct mode) 100 MHz band is used for the link to the satellite, while the satellite to earth link could be in the same band as the direct mode. Coordination is employed to provide a high multiple access capacity and to facilitate the use of satellite processing, providing a high degree of anti-jam protection for the necessarily power limited satellite link.

The satellite will contain a frequency hopping pseudonoise processor. The coordination of the net will be such that only one user is accessing the satellite at any time; the 4.8×10^{-3} duty factor, thus, gives a total of slightly more than 200 accesses with a typical power budget as given in Section 7.3.5.2 of Volume II. The satellite

will have a 10 channel receiver and the frequency hopping part of the processor will accept receptions only from one selected channel at a time. Channel selection is made under control of a pseudonoise generator in the satellite. The same pseudonoise generator provides a 10 MHz code rate for PN processing of the received signal.

The pseudonoise generator in the satellite does not acquire and track the accessing signals. Each access obtains pseudonoise bit synchronism at the satellite by monitoring its own transmitted signal and varying the transmitted sequence generator phase so that time of arrival at the satellite will be accurate within 10-20 nanoseconds. The pseudonoise correlator in the satellite then reduces the signal to narrowband. Subsequent remodulation by the code and frequency hopping to the desired down link channel completes the satellite processing.

This remote mode system has the advantages that:

1. Full AJ processing is afforded each signal;
2. Only one access is made at a time, eliminating power control problems;
3. Satellite processor is simple, requires no active search or demodulation, and operates without satellite control; and
4. The waveform for the remote mode is identical to the waveform employed in the other modes.

The disadvantages are that coordination is forced upon the satellite accesses and that a tracking problem is forced on each of the earth terminals.

4.2.2.6 Landing Mode

The landing mode is assumed to be a beam rider concept so that the burden on the CNI system is essentially that of a communication link. The directivity of the pattern and the local nature of the operation will provide high level signals that are relatively free from interference. The landing mode operation then is considered to be identical to a standard communication channel.

4.2.3 FH/TH HYBRID-COORDINATED WAVEFORM

An alternate candidate hybrid-coordinated waveform may be defined to take advantage of the multiple access performance inherent in a frequency slotted waveform. This FH type waveform will accommodate near/far dynamic range within practical filtering limits. A hybrid TH/FH is described here with parameters selected to meet the requisite data rate and provide a larger multiple access capability than the TH/FH/PN waveform described above. The pulse duration is set at 50 microseconds as sufficient to eliminate intersymbol interference effects. This allows 5000 frequency slots in the 100 MHz bandwidth presumed. The direct mode is noncoordinated.

Use of M'ary sequential decoding* can be considered but is rejected as being excessively complex for CNI application. However, one can consider a block code for the M'ary channel. For example, it is possible to find 64 code words of length 8 digits with M = 8 such that any pair of code words coincide in at most one digit. This code structure has been applied in the TATS modem, currently under development for AFESD and originally suggested by Lincoln Laboratory for a tactical satellite communication channel with multipath.

The demodulator/decoder for this M'ary code is maximum likelihood for the 64 words, with the additional provision that the amplitude from each of the eight frequency slots is peak clipped inherently by an analog/digital converter. The resulting quantized amplitudes are summed over the eight successive digits for each of the 64 words, and the decision is for the code word displaying the maximum sum.

Because the code has rate 0.75 bit/digit, the waveform's duty factor must be set at 1/6 to maintain 2400 bps data rate with the 50 microsecond presumed pulse duration.

The multiple access capacity can be estimated as a first rough approximation by ignoring cancellation effects when an interfering pulse lands on the desired pulse** and assuming peak clipping at the amplitude of the desired signal. Then, an error can be caused only if interfering pulses fall into the pattern of the seven (or eight) different

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* This waveform has been employed in the Lincoln Experimental Terminal, built by Lincoln Laboratory for Satellite Communications.

** This assumption is particularly valid for interference strong compared with the desired signal.

successive frequency slots of some incorrect word. The probability of this is "union" bounded by

$$P_e < 64\epsilon^7 \tag{1}$$

For $P_e = 10^{-5}$, $\epsilon \cong 0.1$, so that the multiple access capacity is computed to be 0.1 (5000) (6) = 3000.

An optimized jamming strategy against this modulation can be postulated as partial occupancy of the channels by tones at the most effective amplitude. Again by invoking the union bound, pair-wise error probabilities can be computed, and for $P_e = 10^{-5}$, the worst-case jamming consists of tones equal in amplitude to the desired signal and occupying, again, about 0.1 of the channels. Thus, the J/S satisfies

$$J/S \cong (5000)\ (6)\ (.1) = 35 \text{ db} \tag{2}$$

for this optimized threat.

The above results on jamming performance shows that the theoretical multiple access performance will be essentially independent of the amplitudes of the interfering signals, and the worst case actually is for equal-amplitude signals. This is true because stronger interfering signals are still* clipped to unity, while the cancellation of the desired signal is diminished. This conclusion, however, is valid only if there is an AGC which can maintain the peak clipping level at the amplitude of the desired signal.

A mechanism for AGC setting is available from the demodulator/decoder by noting that strong interference must be confined to less than 10 percent of the frequency slots if the demodulation is to be successful. Hence, an average of 9 out of 10 slots used for the correct word will not contain strong interference. After each code word is received, the AGC can be derived from the amplitudes of the 8 successive frequency slots of the word deemed correct. Quantitatively, the AGC may be set on the basis of the average power in the correct frequency slot after peak clipping, and this satisfies both AJ and multiple access requirements. For example, the jamming margin corresponds to an average interference per slot 10 db below the desired signal at the 10^{-5} threshold of error probability, so that the presence of jamming changes

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* Obviously, a practical implementation can not accept arbitrarily strong interference without further degrading due to saturation and intermodulation effects.

the AGC setting by at most 0.5 db. This remains valid regardless of the frequency distribution of the jamming.

Now let us examine multiple access to a satellite, assuming satellite parameters as in Section 7.3.5.2 of Volume II (200 watts into a 2 degree beam width satellite antenna). At the earth terminal, the received S/N in 20 KHz bandwidth for one signal is 23 db, and the threshold* is approximately 9 db. Hence, the ideal number of multiple access signals if they were perfectly coordinated to yield ideal power sharing and full time slot occupancy is

$$\text{Ideal Satellite Access} = 25 \times 6 = 150 \tag{3}$$

One way to have ideal power sharing is to provide satellite channelization into 25 separately limited channels, each of 4 MHz width. This, of course, leaves no margin whatsoever, according to the presumed power budget.

If the waveforms are not coordinated, a degradation occurs, as described for a different modulation structure in Section 7.3.5.1 of Volume II. However, the degradation is more difficult to compute than for the PSK or DPSK modulations. To obtain a rough quantitative estimate of multiple access capacity for the M'ary FH/TH waveform, consider only the interference due to the other accessing signals, all assumed much stronger than the desired signal. Then, any digit of the code word being transmitted by the desired signal will be suppressed below receiver noise if another signal occurs in the same satellite channel. Ignoring receiver noise, an error can result only if four or more digits of the code word are suppressed, since two code words differ in at least seven digits and the interference is limited when it appears in the output of any satellite channel.

If four or more digits of the code word are suppressed, an error does not necessarily occur. However, as a somewhat pessimistic computation, we may require that the probability of suppressing four or more digits in a code word not exceed 10^{-5} with random overlap from the interfering signals. The maximum allowed average probability of overlap then is determined to be 0.02, computed from the binomial distribution. Thus, retaining the 25 satellite channels and 1/6 duty factor, the number of interfering signals is allowed to be (.02) (25) (6) = 3, and the multiple access capacity

*An E_b/N_o = 10 db for broadband noise at 10^{-5} error probability is assumed.

is only four signals. This, of course, can be improved by further satellite channelization; for example, with 250 channels, the number of signals is increased to 31. However, each channel still must have an output power capable of overcoming receiver noise, so that the satellite power has been increased by at least* 10 db, just due to the additional channelization.

Thus, it may be concluded that noncoordinated multiple access to the satellite is undesirable for the postulated FH/TH waveform, whereas a respectable multiple access capacity is available when the accessing signals are coordinated to fall into distinct time slots and satellite channels. Hence, a hybrid-coordinated waveform is assumed, with coordination in the remote mode only.

4.2.3.1 Synchronization in Direct Mode

The problem of initial synchronization will be discussed in the context of a sig-signal without data modulation. Thus, a sequence of frequency hop/time hop pulses is presumed with an a priori known pattern. Only the direct mode is of concern since the range uncertainty of 2 milliseconds applies thereto. (In the remote mode, accurate compensation for range to satellite can easily be made.) With 50 microsecond pulse duration, there are 40 time slots to be searched. Sequential detection theory (see Section 7.5.4 of Volume II) shows these can be searched in a serial manner in an average of less than 2 pulses per position; hence, the search time does not exceed $2 \times 40 \times 2 \times 0.3 = 48$ milliseconds, assuming a search step of half the pulse width and noting the average frame of 0.3 millisecond. A further reduction is possible by parallel looks in the vicinity of the pulse, so that the search step size can be more coarse.

An example of a parallel synchronization processor is now described and is conceptually feasible. It depends on the frequency hopper remaining on frequency during the interpulse interval. With this, the reception of a pulse on the selected frequency slot is enabled for the dwell of the hopper. Figure 4-27 shows how the parallel processing is done. The filter output is envelope detected and sampled, say at 20 microsecond sampling intervals starting at the instant the frequency slot was selected. Each sample is quantized, probably to one bit (off-on), and the bit (0 or 1) is added to the previous count for the same delay from initial selection of the frequency slot. This is continued until each of the parallel positions can be dismissed on the basis of its count.

* Some increase is also necessary to allow correct demodulation at the receiver on a partially suppressed code word in the presence of receiver noise.

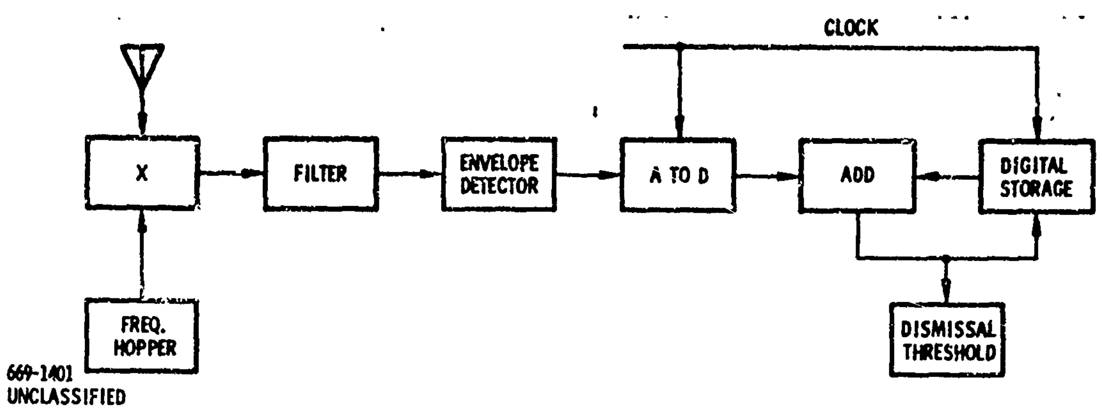

Figure 4-27. Parallel Sync Processor.

As an illustration, if six parallel positions are searched, the synchronization time is reduced by a factor close to six; the actual reduction factor depends on the average time to dismiss multiple positions, compared with the dismissal time for a single position.

An important question is that of AGC prior to synchronization's being detected. One approach is to set AGC on the basis of average power in the frequency slot after peak clipping; this is similar to the procedure discussed above for data demodulation. However, now the AGC setting is based on the response to interference only and may be called a "noise" AGC. For example, if synchronization is to be possible with an average interference power in the selected frequency slot 10 db below the desired signal, the noise AGC will be set the requisite level for the on-off threshold decision until synchronization is detected. Then, the AGC responding to the desired signal amplitude takes over.

4.2.3.2 <u>Tracking and Navigation</u>

A further improvement in tracking accuracy to the order of the reciprocal of total bandwidth is conceptually feasible using the scheme based on coherent frequency hopping and described in Section 4.2.1.10. It will be assumed that the precision tracking still uses only the unmodulated pulses, as discussed above. Although a FH/PN/TH modulation was presumed in the previous discussion of precision tracking, the same concept applies to any FH or FH/TH system, with the observation that practical difficulties increase with the number of frequencies to be synthesized.

For the navigation signals from the satellites, differential time of arrival is to be measured. Thus, the emphasis is on tracking accuracy, and the required data rate (to transmit ephemeris data) is low, roughly 100 bps. For these signals, a greater fraction of the pulses will be unmodulated and allocated to the tracking function. If the ratio is one data pulse to 24 tracking pulses, the required data rate of 100 bps will be met; however, note that the noise threshold and jamming margin still corresponds to 2400 bps, and this can not be improved except by changing the waveform structure to a longer pulse duration. The same general arguments apply to measuring time of arrival of a ground navigation signal is used for ranging in the landing mode.

Synchronization tracking in prior implementations (e.g., TATS Modem) has been applied to this modulation format on the basis of delay-lock tracking on designated unmodulated pulses. A typical ratio is one pulse for tracking and 8 data modulated pulses. This method of tracking should be capable of pulling in to an error of less than ten percent of the pulse width, or about 5 microseconds. This is insufficient for precision navigation.

4.2.3.3 IFF

Up to now, the IFF waveform has tacitly been presumed identical to the COM, and this ignores the lower data rate which suffices for IFF. Assuming 150 bits/sec for the IFF signal, one could simply reduce the duty factor proportionately to approximately 1/100, or an average repetition interval of 5 milliseconds, compared with 300 microseconds for COM. The problem is now that synchronization by a serial search is proportionately slowed and is now unacceptably slow (about one second).

One answer to this difficulty is to adopt a parallel processing scheme for synchronization, similar to that discussed above. Now, the filter output is sampled over the maximum time uncertainty. Since the uncertainty due to unknown range is less than 2 milliseconds, there is no actual serial search required, and the sync time depends on the number of pulses deemed necessary for a reliable decision. If this is approximately 10, the synchronization time for IFF is, as previously, approximately 50 milliseconds.

A further consideration in the IFF function is the interference between several signals interrogating an aircraft all on the same address for the intended receiver. The multiple access capability afforded by the orthogonal time-frequency slots will be

lost if the time of arrival of the signals coincide within 50 microseconds. Only the strongest signal probably can be demodulated correctly in that event, and for equal received powers possibly none of the signals will be received correctly. If the transmitters radiate in absolute time, coincidence of two signals can occur only if the ranges differ by less than 50 microseconds, or 8 n.m. This alone may be operationally adequate.

4.2.3.4 Time Sharing

It is desirable to reduce equipment complexity by time sharing to the maximum extent. With a low duty factor, time sharing of the RF, correlator, synthesizer, and pseudorandom sequence generator is permissible in principle, at the cost of requiring faster hopping in the implementation. Note that with coherent hopping employed for precision range tracking, time sharing of the synthesizer requires the reference to the synthesizer to be phase corrected individually for each signal being tracked, and this is a potential implementation problem, most conveniently handled by the digital phase tracking scheme previously described.

For the FH/TH waveform, a difficulty with time sharing arises from the relatively high duty factor of 1/6. With a multiplicity of signals (NAV, COM, IFF), there is a high degree of time overlap of signals. Let us examine a specific case of requiring four NAV signals with the given modulation format. A procedure for time sharing is as follows: If two or more pulses are going to overlap, choose to receive the first one only. In other words, a given pulse is received if its starting point is not overlapped by a pulse from another signal. The probability of the start of any pulse being clear is $(5/6)^3 = 0.58$, so that there is a loss* of 2.4 db for this degree of time sharing with four signals.

This concept can be extended to consider availability of p processors to receive m signals (m = 4 and p = 1 in the previous example). Now, any pulse can be received if its start is overlapped by p - 1 or fewer pulses from the other signals. The probability of the start being clear in this fashion is

$$P = \sum_{k=0}^{p-1} \binom{m-1}{k} (1/6)^k (5/6)^{m-1-k} \qquad (4)$$

* Overlap in both time and frequency is ignored since it is of low probability. This event introduces a small additional loss.

for signals of 1/6 duty factor. As an illustration, assume a requirement to receive 8 signals. Then, with two processors, the probability of receiving any pulse is

$$P = (5/6)^7 + 7(1/6)(5/6)^6 = 0.67 \tag{5}$$

so that there is a loss* of 1.7 db for time sharing of two processors among eight signals, each of 1/6 duty factor.

Time sharing is not compatible with the parallel processing synchronization scheme previously described which requires the frequency hopper to stay in the selected frequency slot for a large fraction of the interpulse interval. Separate frequency hopper and correlation circuitry must be provided for each of the several signals to which synchronization is desired simultaneously.

Time sharing of transmit and receive functions is also a problem area. In a practical implementation it is not feasible to receive while simultaneously transmitting in a common band. Even if the transmission replaces one of the multiplicity of signals, e.g., COM transmit replaces COM receiver, a further loss from that computed above is entailed for time sharing of processors. This is true because in the receiver, no signal can be received during the 50 microsecond interval of transmission.

4.2.3.5 Satellite Processing for AJ

Since the COM signals for the satellite mode are coordinated, introduction of processing in the repeater becomes conceptually straightforward as a means to enhance AJ. First of all with, say, 25 signals accessing in each time slot, parallel directive beams can be hopped to the locations of the transmitters. This discriminates against jammers outside the beams. The hopping rate for the beams is once per 50 microseconds.

If the accessing signals have a common frequency hopping pattern but are offset by appropriate multiples of 20 KHz into distinct satellite channels, an on-board frequency hopper provides waveform processing to discriminate further against jamming. In fact, with 25 satellite channels of 160 KHz width to accommodate the 8'ary data modulation, the multiple access capability computed above is maintained and fairly good AJ performance is realized.

* In addition to losses already noted, the effect on the error probability due to missing pulses needs to be analyzed, considering the data word structure.

The actual jamming margin is difficult to compute, but a lower bound is the above multiple access computation allowing occupancy of two percent of the frequency channels. Since there are 5000/8 = 625 channels and 6 time slots, the margin is at least 19 db, assuming jamming tones equal in amplitude to the desired signal. A more optimistic computation would allow about ten percent occupancy of the 625 channels, and this implies J/S = 26 db.

4.2.3.6 Basic Transceiver Block Diagram

The basic transceiver block diagram is presented in Figure 4-28 for the FH/TH waveform. Coherent hopping is presumed so that a precision time of arrival measurement is enabled, and a tunable preselector can be applied to discriminate against signals in other frequency slots. This basic block diagram is applied to all three functions, COM, NAV and IFF. As analyzed above ignoring time sharing of transmit and receive, a requirement to receive 8 signals of 1/6 duty factor can be accommodated with modest loss by two RF and synthesizer processors in parallel. Naturally, the decoder must be duplicated for each of the 8 signals, along with the tracking circuitry.

The receive/processor time sharing works conceptually in the following manner. At a given instant, the start of a pulse to be received from one of the signals is assumed to occur. If a processor is free, its frequency synthesizer is set to the requisite slot (and carrier phase) for the selected pulse, and the outputs from the multiple filter bank demodulator are routed to the decoder for the selected signal. If a processor is not free, blank information is fed to the decoder for the selected signal.

The desirability of coordinating the satellite mode has been discussed to improve the multiple access capacity of a power starved repeater. If the orthogonal time-frequency patterns have been assigned,* any transmission is simply advanced in time to arrive in zero time phase at the satellite relay. The receiver is similarly retarded, and the necessary information is available from the navigation computation (to give range to the satellite relay). With this waveform, the timing error for coordination need only be small compared with 50 microseconds, say a few microseconds.

Accurate time of arrival of any signal is measured by deriving a tracking error from the increments of phase from one pulse to the next. As indicated in Figure 4-28, this technique works conceptually as follows. The carrier of each received pulse to be

*The method of assigning addresses is not discussed here.

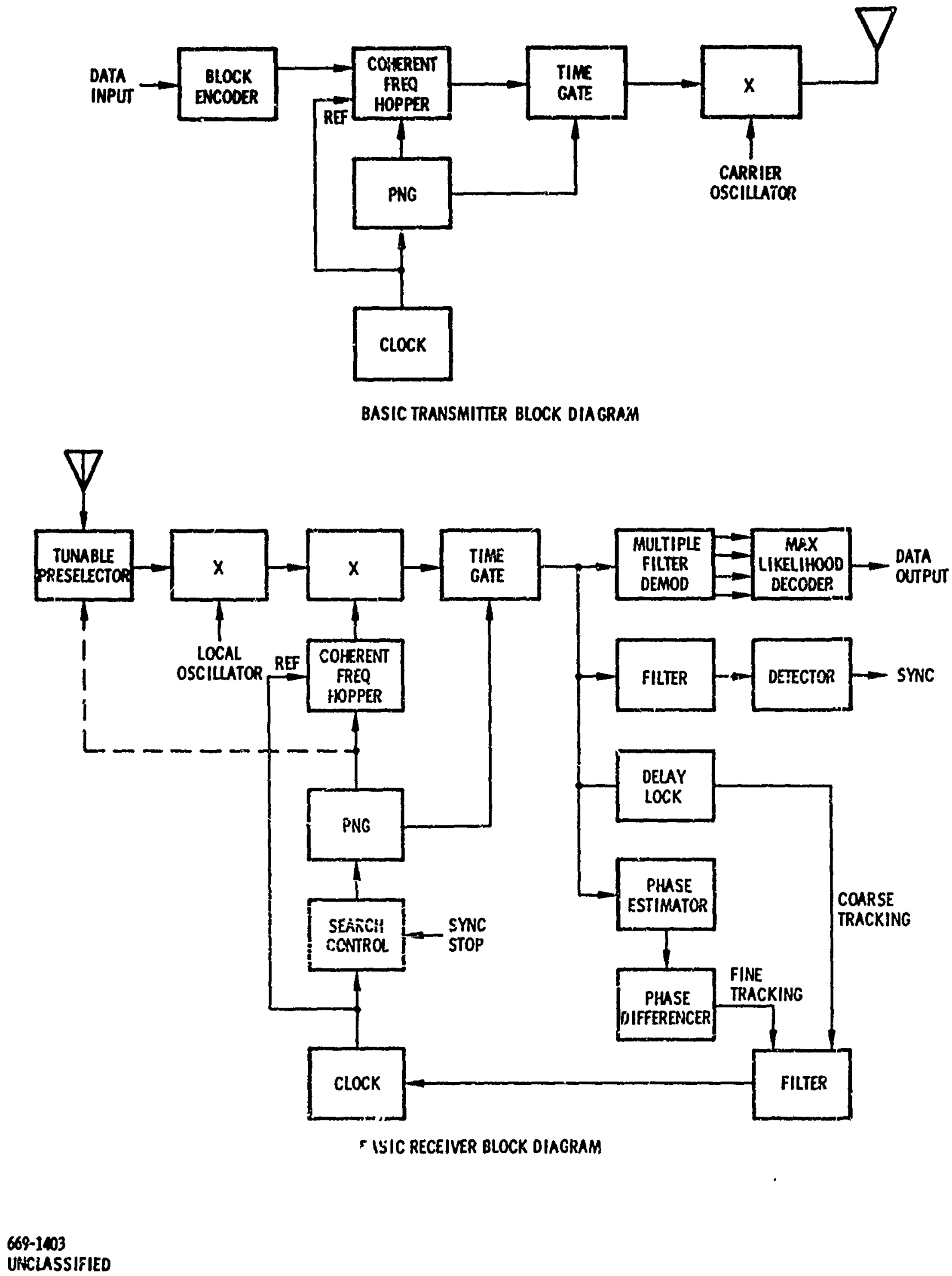

Figure 4-28. Basic Transceiver/Processor Block Diagram.

used for tracking is measured with the coherent synthesizer as a reference. An intergrate-and-dump process yields the in-phase and quadrature components with respect to the synthesizer reference, from which carrier phase may be estimated. These successive phase measurements are differenced as the basis for the error voltage into the fine tracking loop.

The transceiver/processor is sketched grossly in Figure 4-29, following the above concepts. Time shared frequency hoppers and correlation channels are supplied with the necessary digital frequency selection and phase reference, and their outputs routed to the demodulation, synchronization, and tracking circuitry, repeated for each COM, NAV, and IFF signal to be handled in parallel by the processor. The circuitry derives the address code for each signal from a single PNG which contains "code of the day" information and is clocked from the timing standard. The more detailed circuitry can be filled in from Figure 4-28. Note that search is not needed to acquire synchronization to signals from the satellite, if the navigation information is used to compensate for range.

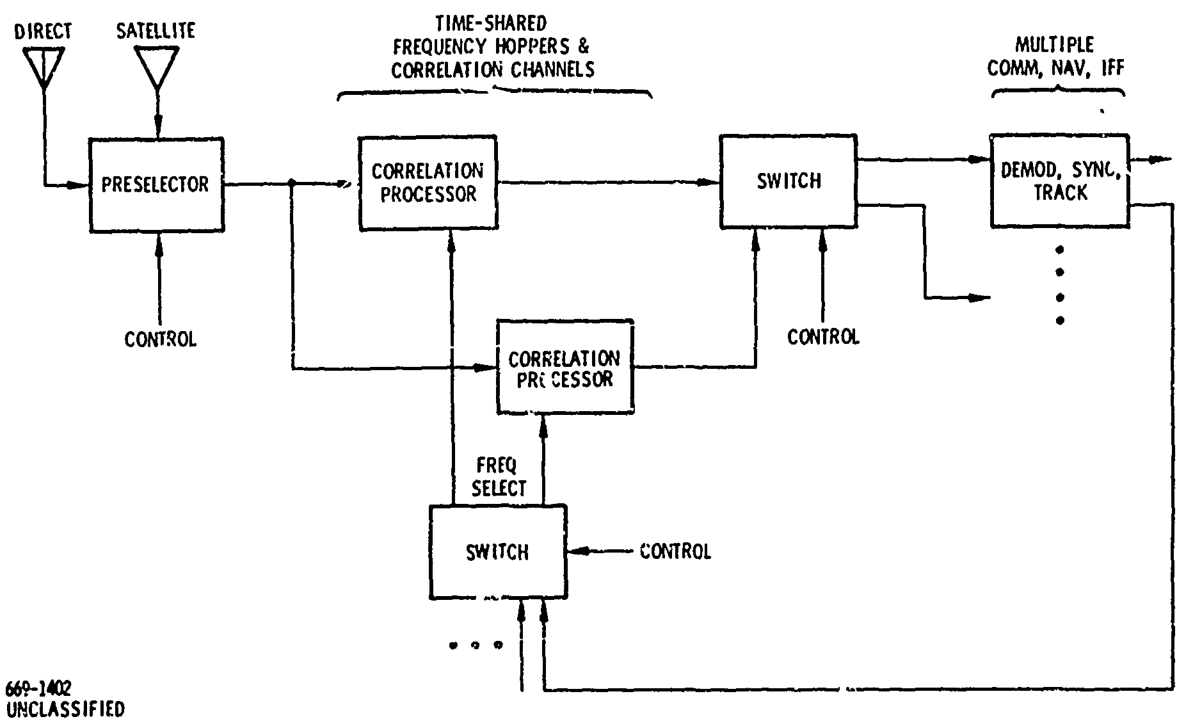

Figure 4-29. Gross Receiver/Processor.

4.3 SELECTION OF PREFERRED WAVEFORM

The rationale is indicated here for selection of the FH/PN/TH coordinated waveform as the preferred waveform concept. Since some of the more specific aspects (details of the waveform) that motivate consideration of this waveform are given in the processor definition discussion, they will not be repeated here, but broader factors concerning waveform type will be discussed here. First of all it should be noted that the distinction in terms of coordinated waveforms is meant to apply primarily to the high density COM traffic after being organized into a coordinated state. The NAV and IFF/signaling traffic, being of relatively low capacity, need not necessarily have their waveforms coordinated (other considerations such as feasibility and implementation govern this case) though the same waveform structure would be used as for COM.

Now a coordinated waveform (for purposes of discussion here) is one whose address parameters, for each signal access, are adaptively adjusted to avoid mutual interference among accesses so that ideally the waveform for each simultaneous signal access becomes orthogonal (different frequency-time slots are used). Furthermore, the rules of waveform parameter adjustment are such that optimum, or more efficient, use tends to be made of any or all of power, bandwidth, and equipment time sharing, which are three of the potential benefits to be realized through use of coordinated waveforms. In fact, the ultimate benefits one could achieve if desired are a truly proportionate share of all satellite power for each access, a bandwidth efficiency of 1 bps/Hz (for binary level modulation) in the direct mode, and a single aircraft transmitter-receiver pair of equipment that will simultaneously serve all the functions of COM, IFF, NAV, and even monitoring of other aircraft transmissions in both of the direct and satellite modes (at least the time sharing for CNI functions is desired though the latter monitoring capability may not be too important).

In contrast, non-coordinated type waveforms offer an essentially unlimited number of available addresses but experience mutual interference among signal accesses on a statistical basis when their slots overlap. They are dependent on powerful coding to minimize this mutual interference (although coordinated waveforms tend not to have such interference and, therefore, no need for coding for this purpose, powerful coding is still desirable to combat jamming, multipath, and to make most efficient use of the satellite downlink power budget). Nevertheless, even with such coding they are

significantly inferior in each category of performance as compared to coordinated waveforms, particularly when nets are postulated to have a high usage factor.

For example, and depending on the specific waveforms considered, a non-coordinated waveform could require 7-18 db more satellite downlink power, would have a bandwidth efficiency of 0.01 to 0.6 bps/Hz (the latter is not feasible if one also wants to have a practically useable amount of AJ so that in this regard a practical upper limit in bandwidth efficiency may be only 0.2 bps/Hz), would have this multiple access capacity degraded when there is simultaneous jamming present (the capacity could be degraded by a factor of 2 with respect to maximum when simultaneous AJ is also degraded by a factor of 3 db with respect to maximum), would have multiple access capacity also degraded by simultaneous presence of multipath(actually in the direct mode one can design against multipath, but in the power-budget-limited satellite mode multipath fading will detract from the power budget available for multiple access), and would have multiple access capacity degraded when equipment time sharing is implemented (that is, additional statistical errors are added when time sharing of a given equipment is added, so that in this sense one must either back off in capacity or else say that one is limited in time sharing capability). As far as dynamic range is concerned one can tend to design for this as one does for coordinated waveforms. However, because mutual interference will still occur statistically, and since the near-far problem represents the worst kind of mutual interference, then in this situation one must also back-off in capacity from the ideal maximum.

The unique system requirements imposed by coordinated waveforms can be summarized as follows:

1. Need to Operate in Common Timing:

Mutually identifiable and distinguishable frequency-time slots must be available for all terminals in each terminal area and at interfaces between terminal areas. The timing precision required for this purpose need be only on the order of 1/10 of a slot width, so that considering plausible waveform designs (2 millisecond slot width) this means only a couple of hundred micro-seconds allowed timing error. This, of course, is not too demanding a requirement. In fact, it is much less precise than the timing required on a single satellite uplink (50 nanoseconds) for AJ processing in the satellite. Each aircraft need not necessarily carry a very stable clock (e.g., atomic)

on board so as to assure this accuracy after extended periods of flight, because a satellite which is presumed visible at all times, could provide this common timing. Alternately, even with no satellite, the timing can be generated on board using the output of the precise NAV channel and ranging information to a given ground or aircraft reference point. Also at interfaces between adjacent terminal areas, and when a common satellite is not visible, timing can be precisely coordinated or corrected by transfer of timing using NAV ranging methods and with a selected terminal acting as a master station. Thus, it is seen that even considering a variety of operational situations, the need to operate in common timing for a coordinated waveform should not be a real problem.

2. Impact of Propagation and Aircraft Motion and Position:

The above discussion has considered only the aspect of airborne time standards with regard to frequency-time slot integrity. Another aspect is that of Doppler shift for high speed airborne terminals. This aspect, however, is not a problem if the size of the slot in frequency is made much greater than the Doppler shift. Plausible waveform designs indicate 100 KHz to 1 MHz for this slot dimension which, therefore, makes Doppler not a significant factor.

Similarly, the variation in propagation delay between dispersed terminals that are within line-of-sight and interference-range of each other, is another aspect of slot integrity. If one makes the guard time in a slot equal to the variation in delay between terminals in adjacent slots then this problem is countered, but at the expense of reduced efficiency from the ideal maximum values cited previously for coordinated waveforms. The reduction equals the guard time divided by the slot time. By optimum allocation of slots to adjacent terminals, via the mechanism of coordination, this ratio can be kept small without requiring long slot times, and thereby long, cumbersome frame times.

3. Limitations on Repeat Jam Invulnerability:

Because of the propagation limitations on slot time, the slots can be permuted only at a low rate, e.g., 500 to 2400 bps. One is therefore

potentially limited in repeat jam invulnerability. Nevertheless, operationally this jam threat may not be important.

4. Need for System Discipline in Use of Slots:

Given that most COM traffic consists of nets, then the particular frequency-time slots for each net would be known a priori to authorized terminals, provided they also have the code of the day and slot timing. No problem in system discipline, therefore, exists in these regards. For situations where the slot allocations are not known to the aircraft terminal desiring to enter a net, then a signalling or order wire link (using non-coordinated frequency-time slots) must be included in the design so that this information can be requested and obtained. Actually with non-coordinated waveform systems, for which the desired net address is not known, they must also have a capability for signalling in this regard to the appropriate terminal. The coordinated and noncoordinated systems are, therefore, no different in that they both require signalling capability for such information to the appropriate master terminal. However, the difference, of course, is that the noncoordinated waveform design needs no special waveform mode for this purpose, whereas the coordinated waveform design does need it; i.e., a special noncoordinated waveform signalling mode but using the same waveform structure. Since the IFF mode has similar requirements, it can be co-used for this signalling function. In this sense no real system limitation or problem exists because of the signalling link with the basically-coordinated waveform systems. Similarly, if and when new slot assignments are to be established on any one or more COM links, the noncoordinated waveform signalling required before one gets into the coordinated waveform state, would be handled by co-use of the IFF mode.

5. Coordination Time:

Coordinated waveform systems require a coordination time, that is over and above the waveform synchronization time. This is primarily for those cases, as discussed in 4., where special signalling is required before one gets into the coordinated state. One must also add to this so

as to obtain the total coordination time, the fact that since each simultaneous signal access is individually slotted, the frame time may be greater than the slot time. As such at the initiation of entry into a net one may have to wait a whole frame time. However, for plausible waveform designs in which interleaved 2 millisecond slots are used, and in which the average frame time is 24 milliseconds, it is found that the coordination time need be only 4 milliseconds to 28 milliseconds depending on whether one seeks to enter at the latter or beginning time of a frame (the coordination time includes 4 milliseconds total for the round trip signalling time). This is reasonably small. However, if multiple signalling tries are required it would be longer. Also more conservative designs for coordinated waveforms (that use more time slots) would have a longer frame time (e.g., 1-1/2 seconds) and, therefore, also have a longer coordination time.

The unique system requirements just discussed for coordinated waveform systems, do not appear to be too difficult to meet. They seem to be justified by the increased system performance, that was discussed earlier, that can be achieved with such systems. On this basis a coordinated waveform design for COM is preferred over totally uncoordinated designs that have been considered. Hybrid designs have also been considered that involve coordinated waveforms in the satellite mode and non-coordinated waveforms in the direct mode. However, it was found that their capacity and efficiency is also rather limited in the direct mode so that coordination in the direct mode as well is still preferred.

4.4 DETAILING OF THE SELECTED WAVEFORM FH/PN/TH - COORDINATED

The purpose of this section is to describe the selected waveform design that results considering all the factors thus far discussed, but also making certain parameter modifications with respect to the candidate waveform design indicated previously in Section 4.2. The modifications involved primarily concern provision of more TDMA slots (so as to increase connectivity, dynamic range performance, and equipment time-sharing capability) on each frequency-time pattern (there are now also fewer parallel frequency-time patterns) and use of wider time slots (so as to increase the guard time against adjacent slot overlap and also to have a proportionate "on-time"). Also refinements are made in the design to make it more generally useful (the timing is compatible

with civilian systems, the waveform parameters are more nearly compatible than before with 621B NAVSAT design and the DSCS satellite design, and the random access mode is designed for more general purpose use). It should be noted, however, that though the modifications improve performance in the areas indicated, they are not achieved without some accompanying disadvantages (significantly longer frame times and buffer storage requirements, long sync times or special sync procedures in cases where because of interference or fading one needs to allow more than one slot for synchronization, and a need for higher peak power transmitters so as to realize a peak-to-average power trade against jamming.)

Relative to the type of waveform description given in Section 4.2, one can describe the modified, basic signal structure as follows in Figure 4-30. It indicates a burst length of 40 microseconds (down only slightly from the previous value of 60 microseconds) so as to minimize the effects of multipath ringing in near-far situations (should the individual bursts be transmitted in a given uncoordinated mode), 36 bits of information and, therefore, 72 encoded digits in a burst (this is much greater than the previous value of 3 bits because with more TDMA slots one must send at a higher rate in each slot), and 8 chips of PN for each digit (this should still be adequate to discriminate against multipath intersymbol interference; the previous value was 10 PN chips) giving, therefore, a channel bandwidth of 14.4 MHz.

Range measurement is performed with this waveform by measuring the clock phase of the 14.4 MHz PN signal for a basic range accuracy of 70 feet; however, by processing one can range to a fraction of this. For even greater range precision (to less than 6 feet) the design here would measure the carrier phase of a sidetone modulation of this waveform, using the coherent frequency hopping implementation concept described previously.

Figure 4-30 then is the basic waveform structure for each frequency-time pattern in which the transmitted frequency is held constant over a 40 microsecond burst length. The design here provides for 10 such frequency-time patterns in parallel, giving, therefore a system bandwidth of 144 MHz.

The basic signal structure of Figure 4-30 is formatted (this means the nature of the sequence of such bursts on a frequency-time pattern) differently for the various modes so as to optimize corresponding performance. Figure 4-31 indicates the

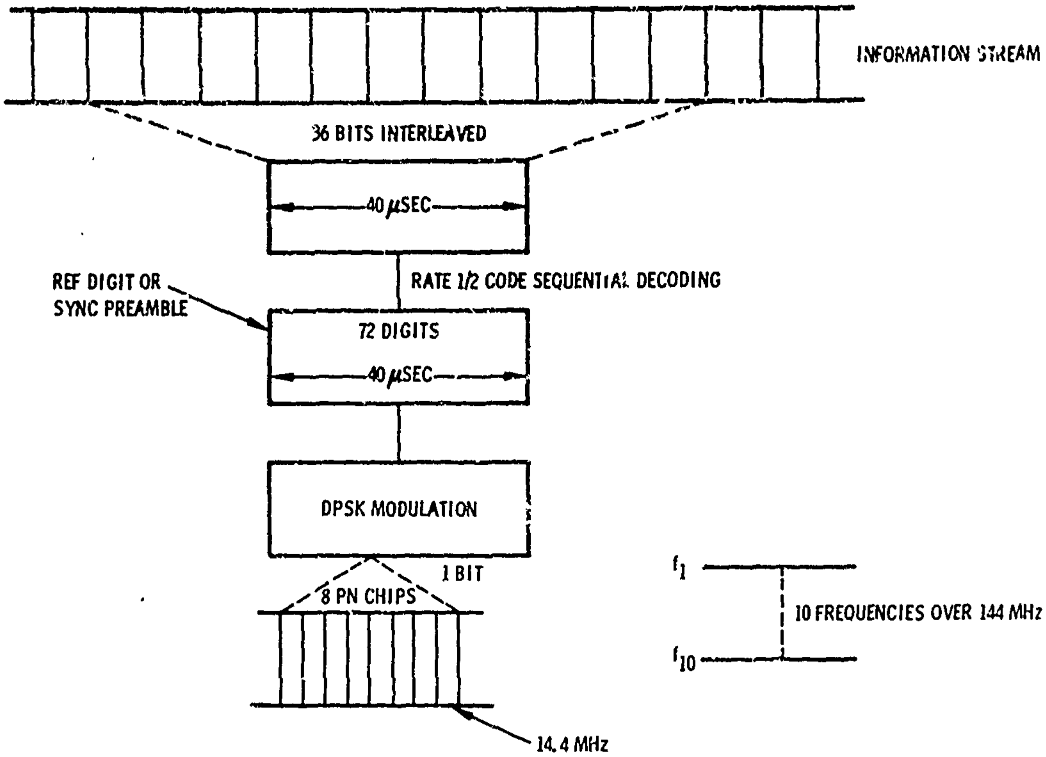

Figure 4-30. Basic Signal Structure

formatting for the direct mode, 2400 bps COM. Each direct mode COM access transmits in an allocated TDMA slot on a frequency-time pattern. The frequency is held constant for the slot duration so that hopping occurs every 6 milliseconds or at 167 hps. The slotting and hopping in each of the parallel f-t (frequency-time) patterns are coordinated with each other. The TDMA slot is 6 milliseconds, of which 4 milliseconds represents "on-time" for the transmission of 100 successive bursts of the basic signal structure, and the remainder is guard time. A total of 3600 bits is sent in a TDMA slot. The first 8 bits, or corresponding 16 digits, constitute a sync preamble whereas the remaining bits are information-bearing. The idea is that under nominal, or Gaussian noise, conditions one should be able to acquire sync, over as much as 1.8 millisecond uncertainty, with this preamble using a digital matched filter (since 8 PN chips constitute one digit the length of the preamble means a 128 bit digital matched filter if one excludes implementation losses). The average false alarm sync probability would be small (10^{-5}) and the probability of sync detection in one try would be about 90 percent. After sync preamble acquisition the receiver should be able to hold sync for the short slot on-time of 4 milliseconds. A new sync acquisition attempt is

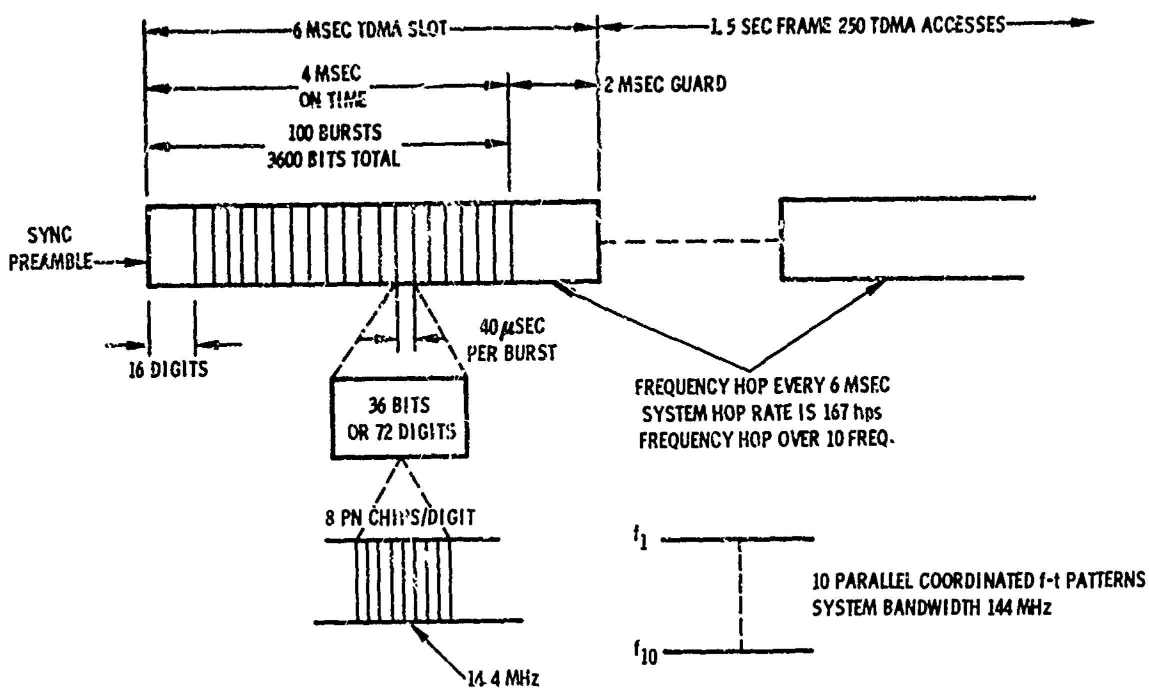

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Figure 4-31. Direct Mode 2400 BPS COM.

made in the slot in each succeeding frame. Linear PN decorrelation is used in the receiver after sync acquisition, with the digital matched filter used only for the latter purpose. The information bit detection error probability after sync acquisition and for the same Gaussian noise conditions is 10^{-5}.

Now for 2400 bps COM there are about 250 TDMA slots in a frame of 1.5 seconds on each frequency-time pattern. System capacity for 10 parallel patterns then is up to 2500 of these 2400 bps accesses. Actually, as will be discussed later, a small portion of this capacity should be reserved for a random access signalling function, for ILS and NAVSAT transmissions, and for incorporating the remote mode in the same frequency band as the direct mode. Also, some of the 2400 bps capacity could be exchanged for equivalent 75 bps surveillance net operations or for higher bit rate data capability.

Now the above, direct mode COM design uses a redundancy of 16 to 1 in a slot (PN plus error correction) achieves a 67 percent slot stacking efficiency, and provides coordinated accesses to up to 2500 f-t slots that are pseudorandomly permuted. The direct mode performance that is obtained because of this can be summarized as follows:

(1) Multipath interference is combatted whether it be in the form of intersymbol interference (both the PN coding and sequential decoding are effective), fading (the sequential decoding and frequency hop diversity are primarily effective), or ringing due to backscatter (the use of coordinated TDMA slots is primarily the factor).

(2) Wide dynamic range operation is possible for large clusters of terminals. That is, since we have 250 TDMA slots on an f-t pattern and since the 2 millisecond slot guard time assures no slot overlap for users in a 300 mile terminal area, then ideally up to 250 terminals in a cluster can operate side-by-side (greater than 100 db dynamic range) without mutual interference. For terminals on parallel f-t patterns, a more modest (e.g., 80 db) but significant dynamic range of operation is possible with respect to each other. (This depends on practical implementation capabilities for frequency division multiple access).

(3) Equipment time sharing (one transceiver for operation in the different modes, nets, and links) is facilitated by the TDMA aspect.

(4) A total processing gain of 48 db is available against external interference such as jamming. This requires that a 24 db peak-to-average power trade (250 KW peak power for one KW average) be realized, however.

(5) As indicated above, a relatively high system capacity (2500 of the direct mode 2400 bps accesses) results even with the simultaneous presence of the different interferences.

(6) The parameters involved have compatibility with civilian system interfaces as discussed in a pertinent information document.* Also as will be discussed later the waveform parameters are compatible with the 621B NAVSAT type of design and with the DSCS Phase II and advanced satellite type of design.

For direct mode operation of groups of accesses at bit rates other than 2400 bps, the typical scheme would be to assign more time slots (for higher bit rates) or

* F. Diamond and A. Kunze, "A Basic Format for Integrated CNI Systems," Rome Air Development Center memorandum.

fewer time slots (lower rates - fewer time slots also mean a longer frame time) in a given time interval, as the case may be. System slotting efficiency is not changed in this way and the implementation is simple. However, plausible exceptions to this procedure are as follows:

(1) For non-vocoded voice; e.g., 19.2 kbps delta-modulated voice, the higher redundancy inherent in the digitized voice signal would allow operation at a higher channel DPSK bit error rate. As described in Section 7.3.3 of Volume II, data obtained by Magnavox in tests at Ft. Huachuca indicate a 10 percent bit error rate gives better than 90 percent speech intelligibility. The rate 1/2 sequential decoding (a 5 percent bit error rate in gives 10^{-5} error rate out) should, therefore, be deleted for this voice application, meaning then that one needs to preempt 8 times the number of time slots rather than 16 times as compared to 2400 bps vocoded voice. The system direct mode capacity would then be 312 delta-modulated voice accesses as compared to 2500 vocoded (2400 bps) voice accesses. Because multipath intersymbol interference can be of equal strength, or greater, as compared to the direct path, one would not seek to eliminate the PN coding in addition so as to economize further on time slots (the bit error rate could then approach 0.5, an obviously intolerable value).

(2) For low bit rate data; e.g., 75 bps, if one assigned 32 times fewer time slots per access, the frame time would be correspondingly longer; i.e., 32 x 1.5 = 48 seconds. The system direct mode capacity could be up to 80,000 of these low bit rate accesses.

Though the FH/PN/TH waveform design here is basically a slot-coordinated one, it appears desirable to include in the direct mode a limited capacity for uncoordinated waveform transmission (but, of course, using the same basic signal structure). Examples of plausible use for such a capability include:

(1) Low bit rate signalling to obtain slot assignments given that one must provide more addresses than there are slots, or that an aircraft in special circumstances just does not have the slotting information.

(2) Random access type of IFF for situations where it is operationally not desired or not possible to have all aircraft in coordinated slots - also for this random access type of IFF one may want to form each aircraft instantaneous address in accordance with his known instantaneous NAV position so that because of the large number of possible addresses one has to use uncoordinated slot transmissions.

(3) Transfer of precise timing corresponding to an accuracy much better than the unknown propagation delay of the given aircraft to another terminal - this will be for special situations such as where the aircraft clock has failed or there is a long interval between updates. The transfer of timing process involves transponder ranging by the aircraft using uncoordinated slots.

(4) Emergency or special sync acquisition mode so as to obtain sync in a short time. This would be for special interference situations (such as intelligent jamming or fading) where one cannot reliably acquire sync in the normal way using the sync preamble in one TDMA slot, so that multiple tries in each successive 1.5 second frame would otherwise have to be made. The use of uncoordinated slots for random access sync would allow multiple independent slot tries to be made in a short time. After acquiring initial sync this way, one can plausibly switch to the slot-coordinated TDMA mode and maintain sync track in this mode.

The formatting for the random access mode involves sending individual 40 microsecond bursts in pseudorandomly selected burst intervals (rather than sending a continuous sequence of them) and hopping the frequency from burst-to-burst. The idea is to randomize the burst slot selection in both frequency and time for each burst and to design for a capacity corresponding to fractional occupancy of all the slots. All accesses in this mode are uncoordinated in their burst-slot selection. Repetition coding (which is also combined with error correction coding for transmission of significant data) then protects against burst errors such as caused by strong multiple access interference or fading. The PN coding in a burst is not effective against the very strong interferers (10 db or more stronger). However, for the average case where half the interferers are weaker, it aids in doubling capacity over the worst case situation.

Nevertheless, the PN coding discriminates against other types of interference within a burst; in particular against multipath intersymbol interference.

The waveform design here would multiplex the random access burst transmissions in 6 millisecond TDMA slots (using all parallel f-t patterns for that slot time) that are exclusive for this purpose but that are interleaved with the TDMA slots for the main coordinated COM waveform transmissions. Figure 4-32 indicates the multiplexing and formatting of the random access mode that is arbitrarily indicated as using every tenth TDMA slot. Now in a TDMA slot "on-time" of 4 milliseconds, there are 100 of the 40 microsecond burst slot intervals available, and with 10 parallel f-t patterns one then has a total of 1000 burst f-t intervals. A plausible system design would allow for an average occupancy of all slots of 10 percent so that the multiple access interference probability for the worst case of all strong interferers would also be 10 percent. For the case of random access data transmission where error correction coding is used (and bits are interleaved over subsequent bursts) the slot interference probability of 10 percent still allows a 10^{-5} data bit error rate out. For the case of random access sync acquisition attempts, the design for a 10 percent slot hit probability means that it takes only two repetitions to get a 99 percent success probability and a corresponding number of additional repetitions for further reliability, all of which seems like a plausible design. Similarly, the case of random access ranging transmissions and of low-bit data (one burst of 36 bits could be a whole message) transmissions (such as for random access IFF) the 10 percent probability means that only a few repetitions are needed for a reasonable design probability of success. Now the interpretation of the 10 percent slot occupancy in terms of the number of simultaneous, independent accesses, of course, depends on the length of the messages being handled and on the design probability of success that is desired. As an illustration, it can allow for 10 percent of all the 2500 direct mode COM accesses in a worst case interference environment to make a random access sync acquisition attempt in a 1.5 second frame, and with 10 repeated transmissions made for each sync so as to achieve high reliability. Alternately, the random access mode can accommodate up to 250 separate, random access, IFF queries in a worst case interference environment and in a 1.5 second interval with 10 repetitions of each query made for increased reliability. As a further alternate example, the random access mode, with the parameters indicated in Figure 4-32, would allow 25 random, 2400 bps COM accesses (in this case the bursts are not repeated but bit interleaving and error correction coding are used to achieve a 10^{-5} bit error rate) in a worst case

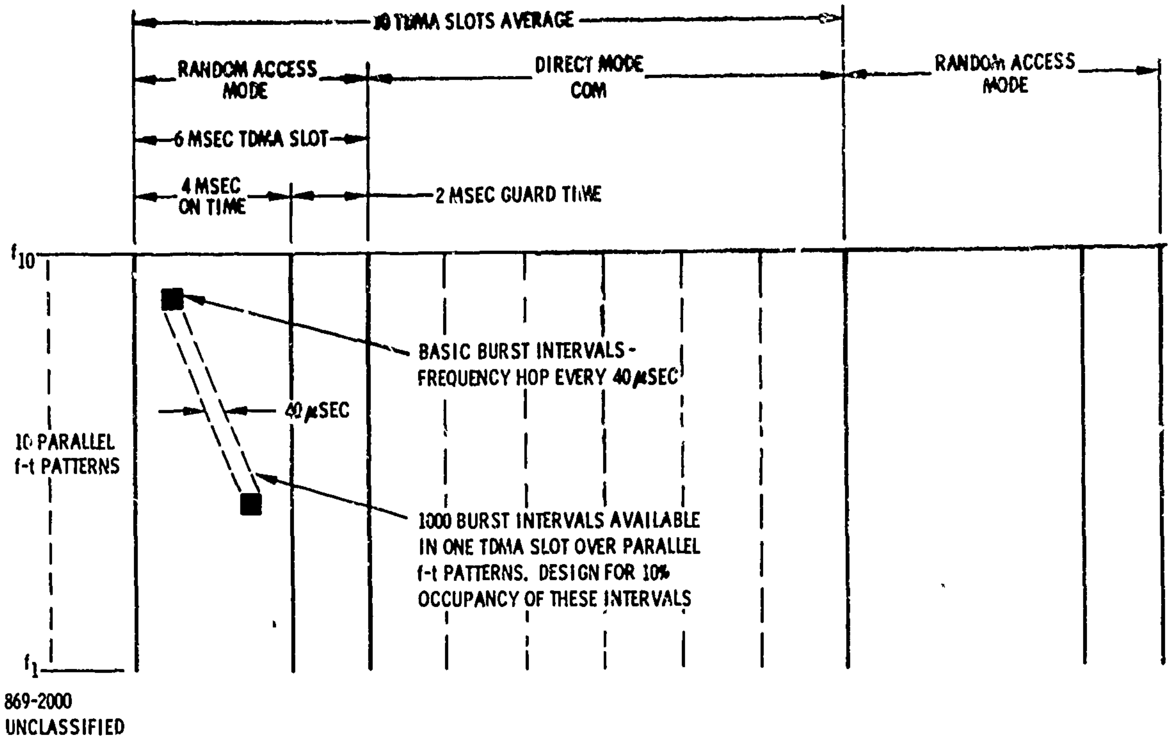

Figure 4-32. Random Access Operation with Respect to the Direct Mode COM Traffic.

interference environment. As indicated previously this capacity is doubled for the average case interference environment.

For the ILS mode, the referenced information document* indicates that a civilian compatible design could be one that uses a separate parallel f-t pattern, and that 30 millisecond intervals are allowed on a TDMA basis for az broadcast, el broadcast (both broadcasts are by a scanning beam ground transmitter), and active ranging by the aircraft. For the waveform design here, the az and el broadcasts would use a 40 microsecond burst (as per Figure 4-30) for each transmission (done every 150 microseconds) and employ position modulation of the burst (a deviation of 40 to 120 microseconds) to convey angular pointing information (as indicated in the referenced information). Also, to avoid the need for stable clocks a reference burst, in addition to the position-modulated burst, must be transmitted each time. Now though our sequence length of 40 microseconds (needed only to be common with our waveform for the rest of CNI) is much longer than the pulse widths for the corresponding civilian

* F. Diamond and A. Kunze, op. cit.

ILS system for az-el broadcasts, the PN coding of our bits allow us to resolve a position-modulated burst with respect to the reference burst down to one chip width or even less. For 14.4 MHz PN bandwidth this is better than 0.07 microseconds. This position measurement capability then is compatible with an ILS that represents 1 degree of motion in 8 microseconds, and, therefore, represents the required angular accuracy of 0.05 degrees by 0.4 microseconds. The 6 to 1 greater capability our waveform has, can be taken advantage of either by an ILS antenna with greater coverage or by a system requiring greater angular accuracy than 0.05 degrees.

For active ranging in 30 milliseconds by all aircraft in a 20 mile landing zone (the RTCA data on ILS described in Section 7.7.2 of Volume II indicates in worst cases that up to 50 aircraft may be there) TDMA of all aircraft ranging transmissions is used. Also, the ILS transponder replies on the same frequency it receives on but with one interval of delay. This latter $t_1 - t_2$ concept means demodulation and remodulation in the transponder (since we wish to avoid an RF delay line implementation). An alternative $f_1 - f_2$ concept, for which linear frequency translation is used and no demodulation is required, is not feasible here given that ILS will be in the same frequency band as the rest of CNI (with just 10 parallel frequency-time patterns an inadequate channel separation is available to make the $f_1 - f_2$ concept feasible). Figure 4-33 indicates the waveform formatting for ILS ranging with the TDMA concept here. As seen a single 40 microsecond burst is transmitted by each aircraft in his TDMA slot, which is 300 microseconds wide. The guard time of 260 microseconds which exists in each slot is adequate to avoid adjacent slot overlap interference between a transponded replay and an interrogation for aircraft that are 20 miles out in the landing zone. Also, the slot size accommodates ranging by 50 aircraft in 30 milliseconds, as indicated in Figure 4-33 with a new range fix (and also az-el fix) being obtained every 0.18 seconds. Since 36 bits of information can also be transmitted with each burst the scheme here also provides a data link of 200 bps. The single f-t pattern which is used for the ILS function is hopped at the same hop rate as, and in coordination with, the rest of CNI. The permutation through 10 f-t patterns at least gives some protection against burst errors.

Now transponder ranging with the basic waveform structure of Figure 4-30, and using just a measurement on the PN clock phase, gives an accuracy better than 70 feet (a PN chip width). One, of course, would already have this accuracy from the NAVSAT system upon entering the landing zone. To achieve the required, increased accuracy of 6 feet, two alternatives are plausible with the waveform structure here. One can

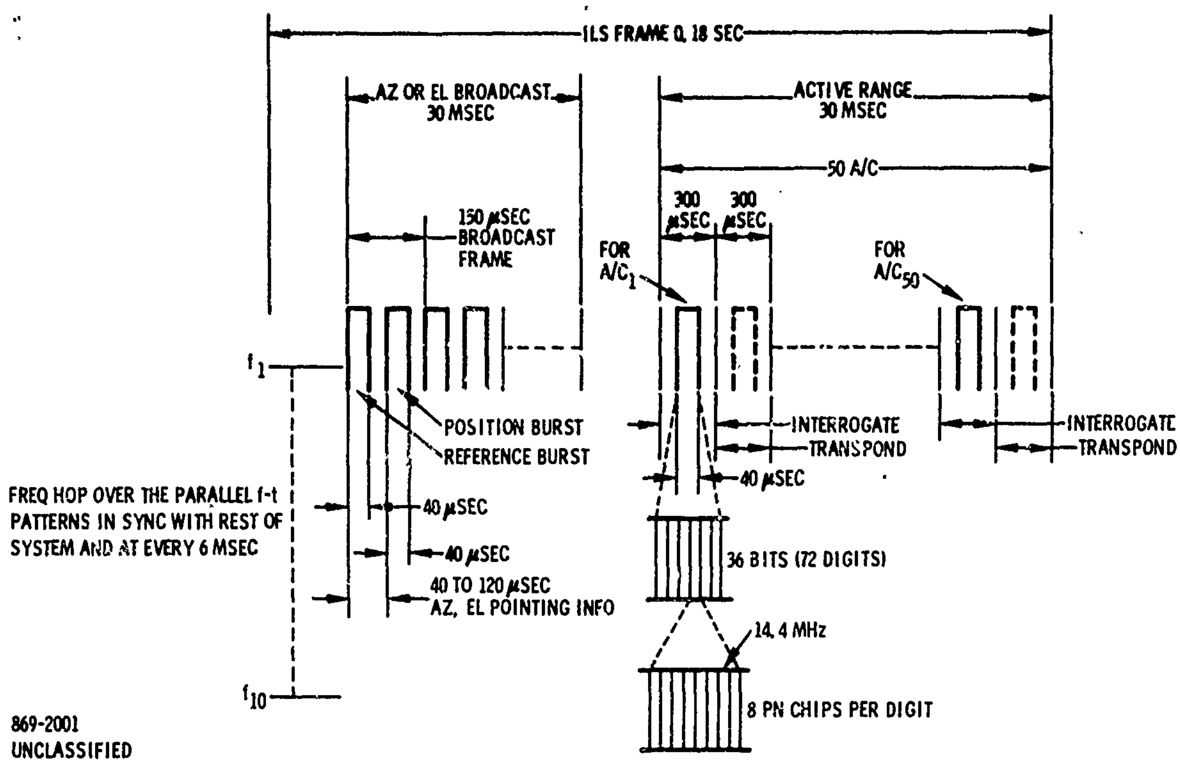

Figure 4-33. ILS Active Ranging and Beam Rider Waveform as Seen at the Transponder.

depend on a strong enough transmitter to give a 21 db S/N and, thus allow one to measure range theoretically to 1/12 of a PN chip width. Alternately, a 100 MHz sidetone modulation of our basic waveform structure is used, and the carrier phase of the demodulated sidetone is used. For the parameters of the ILS design here, where one would range on a single 40 microsecond burst of sidetone, DSBSC modulation by the sidetone (rather than the SSBSC discussed in Section 4.2.1.10) can be employed with tone extraction for the range measurement. This would require then that actually two parallel f-t patterns at the edges of the band (though they are 50 MHz from the reference frequency they are spaced by 100 MHz) be used for ILS. The SSBSC described previously would compare phase over pulses separated by 0.18 second, and this may be an excessive separation for practical use.

As mentioned above, demodulation and remodulation is performed at the ILS transponder. It comprises frequency dehopping, PN decorrelation, sidetone demodulation and phase memory in a phase-lock loop, and then corresponding remodulation of

the sidetone, PN re-correlation, and frequency re-hopping. The implementation is indicated in Figure 4-37. It is actually similar to that for satellite processing, as in Figure 4-38 except that the latter does not perform sidetone demodulation and remodulation. Also as seen from Figure 4-37 the implementation is not really complex. The reason is that all aircraft already have NAVSAT information so that they are already in sync with the PN generator and frequency hopper in the transponder. Therefore, only one of the latter is required to process all the aircraft. Also, the same generator and hopper can be used for the remodulation process at the transponder.

For remote mode COM the formatting concept is the same as that in Section 4.2.1 except that the modified waveform here will have the satellite accesses interleaved on a 40 microsecond burst basis rather than a 60 microsecond one. Also, 250 accesses are TDMA-combined on a single f-t pattern in a 15 millisecond frame interval. The frequency hop interval is still 6 milliseconds, however, as in the direct mode. The basic formatting concept involved, and which has not been changed here, is that since propagation delay on each link to the satellite can be predicted accurately one does not need any significant guard time in adjacent TDMA slots, and, therefore, these can be made smaller, such as equal to the burst interval length. What one gains because of this is that:

(1) Since the frequency on-time of the signal transmitted from any one access is much shorter (150 times less) less opportunity exists for an intelligent uplink jammer to intercept and analyze the instantaneous frequency in use so that he can concentrate his power on that one frequency.

(2) The frame time for each signal access is also much shorter now (100 times less) so that one can more easily design for interleaving of bits over successive frames to give diversity against partial slot jamming, multipath fading, and sync burst errors. That is, the buffer storage problem is simplified in this regard (though one cannot do this in the direct mode, presumably these factors are less likely to be a problem so that one does not have to do anything to combat them).

(3) With a 40 microsecond on-time, the ability to make a peak-to-average power trade (and, thus, gain AJ processing gain) is

facilitated, as compared to the case where one has an on-time of 4 milliseconds.

(4) In the case of sync acquisition (special cases of sync reacquisition are what are meant here in the satellite mode) the total time it takes, including repeated attempts for the case of severe interference, will be shorter since the frame time is shorter.

Figure 4-34 indicates the signal formatting for remote mode COM as just discussed.

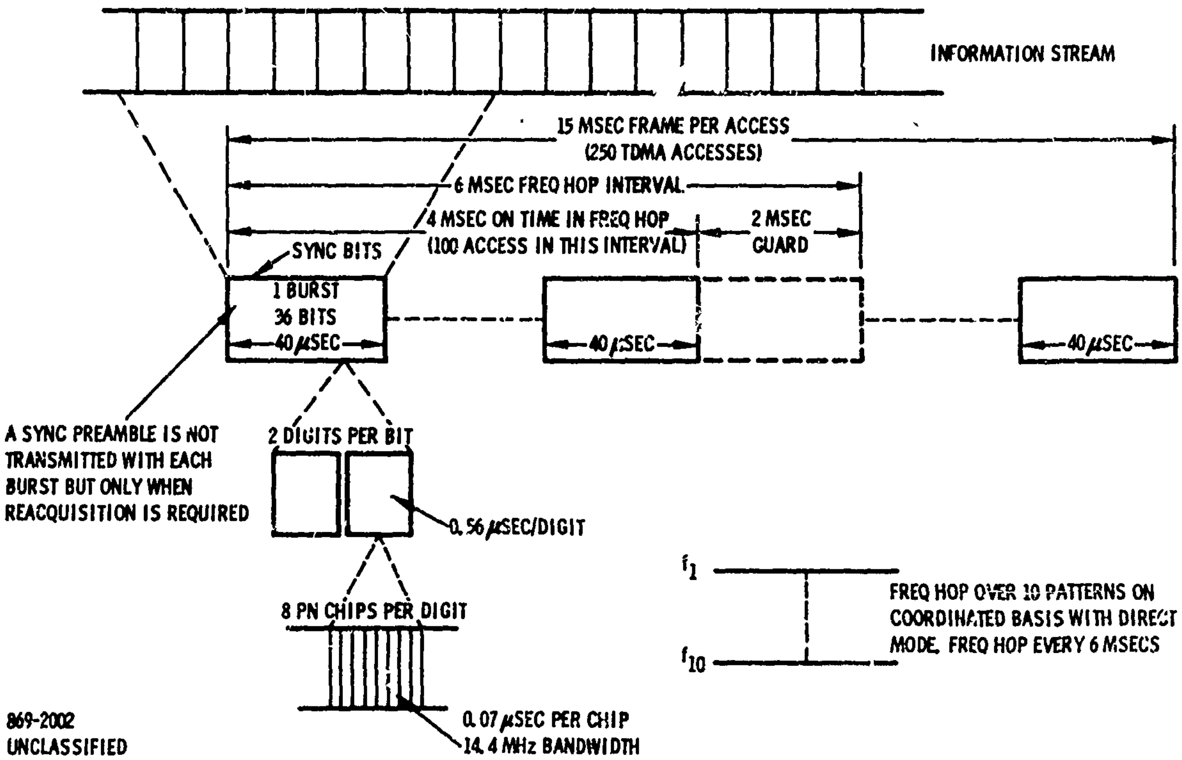

Figure 4-34. Remote Mode 2400 BPS COM.

With the modified parameters of the current waveform design, new advantages in satellite COM result over the design in Section 4.2.1. Namely, since all accesses (up to 250) can be accommodated on a TDMA basis on a single carrier, we gain all the added advantages of this type of multiplexing:

(1) Each access obtains full use of satellite power in its time slot and there are no intermodulation and power control problems. All these are of advantage for the case of a power-starved satellite downlink.

(2) Since we have no power control problems and since a single f-t pattern is involved the satellite can consist of a straight-through, hard-limited repeater in the clear mode.

(3) In the case of uplink jamming, for the case of a satellite which is power-rather than bandwidth-limited, the processing one would provide in the satellite (frequency dehopping and PN decorrelation; see Figure 4-38) need involve only one IF channel rather than 10 as previously.

(4) For power-starved downlinks a single 2 degree satellite antenna beam can be time-hopped and shared among the different accesses requiring it (it is noted that though connectivity is denied to these small terminals by use of a narrow beam they probably do not need the connectivity anyway for normal COM. For other terminals that need the connectivity and have directive antennas, the set of all satellite downlink signals can be sent redundantly in parallel on an earth coverage, broadcast beam).

(5) Similarly to discriminate via narrow antenna beams against uplink jamming, a single narrow satellite beam can be time-hopped and time shared among the different small aircraft accesses.

(6) A TDMA-type of signal format here makes the waveform more nearly compatible with DSCS Phase II and advanced satellite design concepts.

The remote mode COM f-t pattern just discussed for this waveform design concept, is in the same frequency band as, and in parallel with, the f-t patterns in use by the direct mode COM. This is the case for the satellite downlink whereas the satellite uplink would be in a separate frequency band. The satellite downlink f-t pattern is, however, hopped in a coordinated basis with the direct mode traffic over the 10 parallel f-t patterns. The reasons for multiplexing the remote mode COM on this FDMA basis, rather than say on a TDMA basis with the direct mode are:

(1) By making the satellite "on-time" designed for almost 100 percent duty cycle (whereas a TDMA basis would have a 33 percent duty factor), efficient use is made of satellite power.

(2) As discussed previously, short frame times can be designed, with all the attendant advantages indicated, whereas a TDMA basis would make the satellite frame times the same 1.5 seconds as the direct modes.

As for dynamic range problems with regard to receipt of the satellite downlink signal, when there are adjacent, strong direct mode transmitters, coordinated allocation of slots should help the situation. After all, the typical aircraft terminal does not have to receive on all slots, and there are 250 of them on an f-t pattern.

For remote mode NAV a separate broadcast f-t pattern, that is in the same frequency band as the direct mode, is used by all 4 NAVSAT's (a 621B configuration is assumed) on an uncoordinated basis with respect to each other. The satellites' broadcast ranging signal continually and of the same signal structure as in Figure 4-30. The information modulation of 36 bits on each burst is sent repetitively. Ranging is done solely by measuring the clock phase of the PN in each burst transmission. Sidetone ranging is not used in the NAVSAT mode. Each NAVSAT transmission is processed by the receiver sequentially and for an equal share of the time in each 1 second fix interval that the receiver is not programmed to process COM or IFF signals (one time shares one receiver for the different functions). A 100 Hz bandwidth NAV receiver processing is assumed (a 10 millisecond continuous portion of the NAVSAT ranging signal is processed) to get an E/N_0 of 7 db (sufficiently positive to get in the linear region of the phase lock loops, and also adequate value for detection of error correction coded bits to 10^{-5} error rate). The range measurement possible with a 14.4 MHz PN signal in this 10 millisecond look interval is 30 feet (to less than 1/2 a PN chip interval). A total of 25 such repeated measurements of range to each NAVSAT are available in the portion of a range fix interval that the receiver is tuned to each NAVSAT. Digital processing of these multiple measurements will improve the ranging accuracy of 30 feet indicated for just ideal Gaussian noise (ideally 6 feet is achievable, but because of the digital nature of the processing, the results will not be this good). Also, for other interference environments, such as multipath fading, the digital processing of the multiple measurements, which each experience independent fades will enhance performance. Furthermore, it should be noted that the use of a 100 Hz bandwidth for each

NAV measurement (and allowing the multiple looks possible in an extended measurement time to be processed digitally rather than before detection) allows the track loop to hold lock under severe range accelerations without the aid of an inertial platform. Figure 4-35 indicates the NAVSAT signal structure as just described.

Ideally, one would hop the NAVSAT f-t pattern over the 10 parallel ones so as to get a frequency diversity against various types of interference. However, for the design here, this would require overlaying NAVSAT signals on the direct mode on an uncoordinated basis, or else coordinating the hopping with the direct mode on a very slot basis; e.g., every few hundred milliseconds. In the former case one has only the 2 millisecond slot guard times (in each TDMA slot) as the unused portion of system slotting to get the NAVSAT signals through (one cannot just process out the in-slot interference by the direct mode since in worst cases it can be over 100 db stronger), plus one also has the random access slots which are only 1/10 of all slots. This would all require almost 5 db more NAVSAT power budget against Gaussian noise and also the ability to amplitude-discriminate against direct mode interference (since one does not

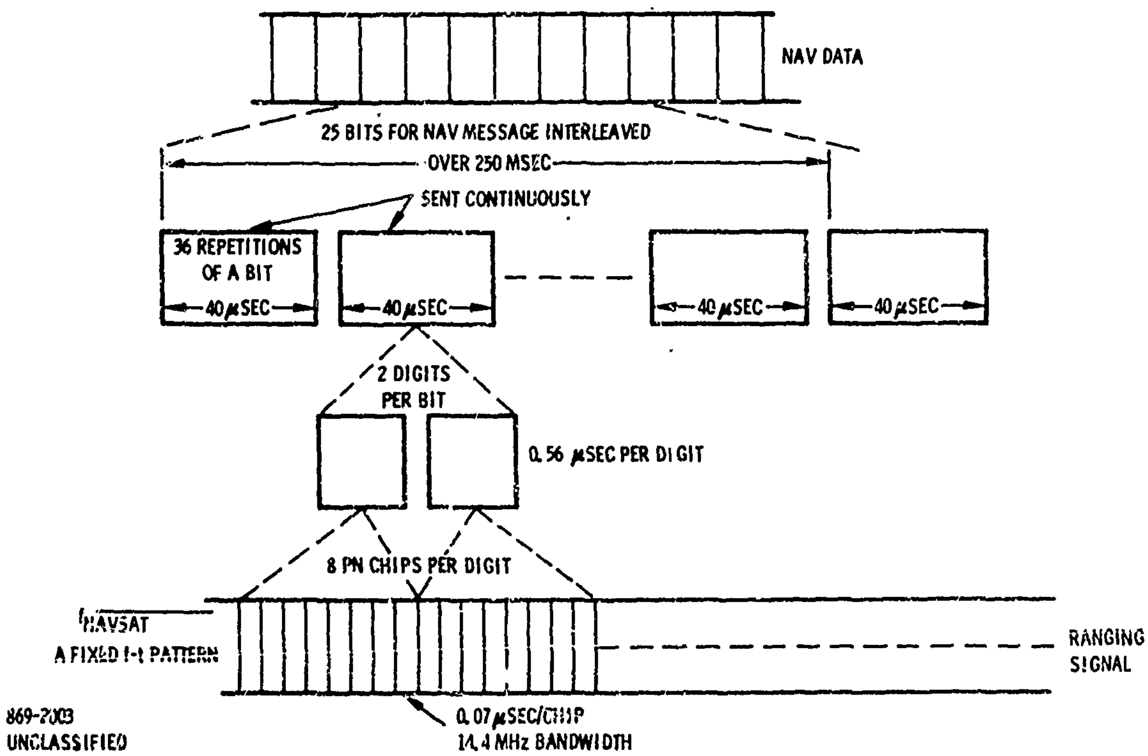

Figure 4-35. Transmitted Signal Format for Remote Mode Navigation.

know where in a slot the 2 millisecond off-time will occur owing to propagation delay uncertainties). In the latter case the slow rate of hopping coordination allowed may not give any frequency diversity in the 1/4 second or so of time in which a receiver is tuned to a NAVSAT. The former, therefore, would seem to be a more promising design. However, for simplicity reasons the waveform design in this section assumes that the NAVSAT frequency-time pattern is fixed (i.e., no hopping, at least on a daily basis). Now, though the 4 NAVSAT's co-use a frequency-time pattern, their mutual interference to each other has a dynamic range of only 6 db or so (plus antenna pattern variations). It is easily processed out in the receiver.

Figure 4-36 indicates the multiplexing of all the functions that have been discussed in one 144 MHz frequency band, with the exception that the satellite uplinks are in a separate frequency band and are not shown (just the COMSAT and NAVSAT downlinks are indicated in Figure 4-36). Indicated is the situation where the NAVSAT's are on a fixed f-t pattern (not hopped with the rest of CNI), and also where ILS is on a single fixed f-t pattern. Because it is not difficult, the option exists of actually hopping the ILS f-t pattern in coordination with the rest of CNI. Also, if coherent sidetone ranging, together with DSBSC modulation of the sidetone, are implemented then 2 f-t patterns, rather than one, would be needed for ILS. Now all direct mode COM, and the satellite downlink COM do have pseudorandomly hopped f-t patterns that are permuted with each other in a coordinated or orthogonal sense. A single pattern is indicated for the satellite COM whereas just 7 f-t patterns are indicated for direct mode COM (the previous discussion in terms of 10 patterns or 2500 accesses was for maximum direct mode capacity; with the multiplexing of the other CNI functions in the same band one must now derate this capacity for the direct mode COM). The patterns involve permutation of frequencies for each 6 millisecond slot interval (frequency hopping for COM, therefore, is at 167 hps), and also permutation of time slots in each frame for TDMA accesses on the same f-t pattern. It is noted once again that the frames are of different length for the direct and satellite modes (1.5 seconds and 15 milliseconds, respectively), and, therefore, the time slot permutation is done at the corresponding different frame rates. The permutation of time slot and frequency is shown in Figure 4-36 for just one illustrative direct mode f-t pattern. Also, indicated is the permutation of just frequency for the total set of satellite COM accesses on the satellite f-t pattern. It is noted that the time slots are also permuted for each satellite access. This is not shown on Figure 4-36 because, since the satellite frame is much smaller

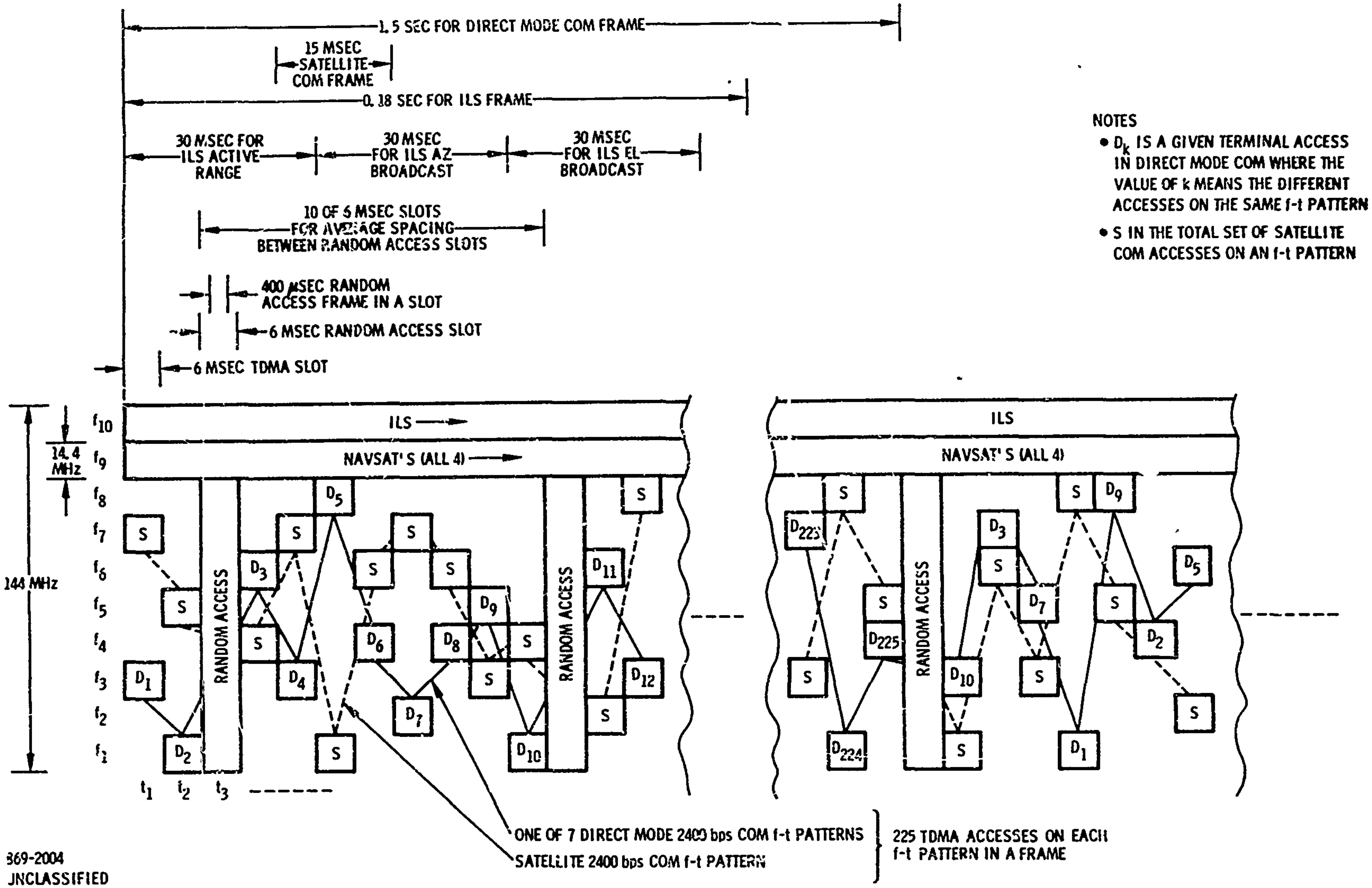

Figure 4-36. Multiplexing of CNI Functions.

(15 milliseconds) the scale would be much smaller. Now for each 2400 bps COM pattern, a total of 225 TDMA accesses are indicated though previous discussions were in terms of 250. That was for maximum capacity but since we now interleave a 10 percent random access capability, indicated in Figure 4-36, the COM capacity must be lowered correspondingly. A total of 1575 direct mode 2400 bps COM accesses thus result, whereas the satellite mode with a single f-t pattern has a total of 225 COM accesses. The random access capability indicated in Figure 4-36 uses all 8 parallel f-t patterns in each of the 6 millisecond slots allocated to it. There is a permutation in the slot of frequency every 40 microseconds, and of the subtime slot (or 40 microseconds) in every subframe interval (depending on the use one makes of the random access function the frame could be as small as the 40 microsecond subtime slot interval or typically its average value would be 400 microseconds). The permutations of the different random accesses are, of course, uncoordinated with each other. The total 6 millisecond time slot allocated for random access is, of course, coordinated with the rest of COM in that exclusive slots are assigned and they contain a 2 millisecond guard time. The average spacing of the allocated random access time slots is 10 of the 6 millisecond slots. It is noted that the actual spacing is also pseudorandomly varied so that 10 is indeed an average value.

Let us consider now the gross functional diagrams for the modified waveform design here. The transceiver diagrams (both transmitter and receiver) indicated in Section 4.2.1 for the previous design are still essentially applicable here, and therefore they are not redrawn (see Figures 4-19 and 4-20). It is noted, however, that the coherent frequency synthesizer and sidetone ranging functions indicated in the transceiver diagrams are no longer needed in the NAVSAT mode. Also, it is noted that since the implementation intends to time share one of a kind of each piece of equipment, then with the increased use of TDMA in the current modified design, this is facilitated even more. Figure 4-37 indicates the ILS transponder processing. Now, relative to satellite waveform processing, Figure 4-38 here gives the current modified description. Basically, it indicates that only for uplink jamming and for a power-limited satellite (rather than a bandwidth-limited one), would one include waveform processing in the satellite. Otherwise in the clear mode the satellite functions as a straight-through, hardlimited repeater (since all accesses are TDMA multiplexed on one f-t pattern and power control problems therefore no longer exist). In the jam mode frequency dehopping and PN decorrelation is indicated together with a single, IF-limiter combination

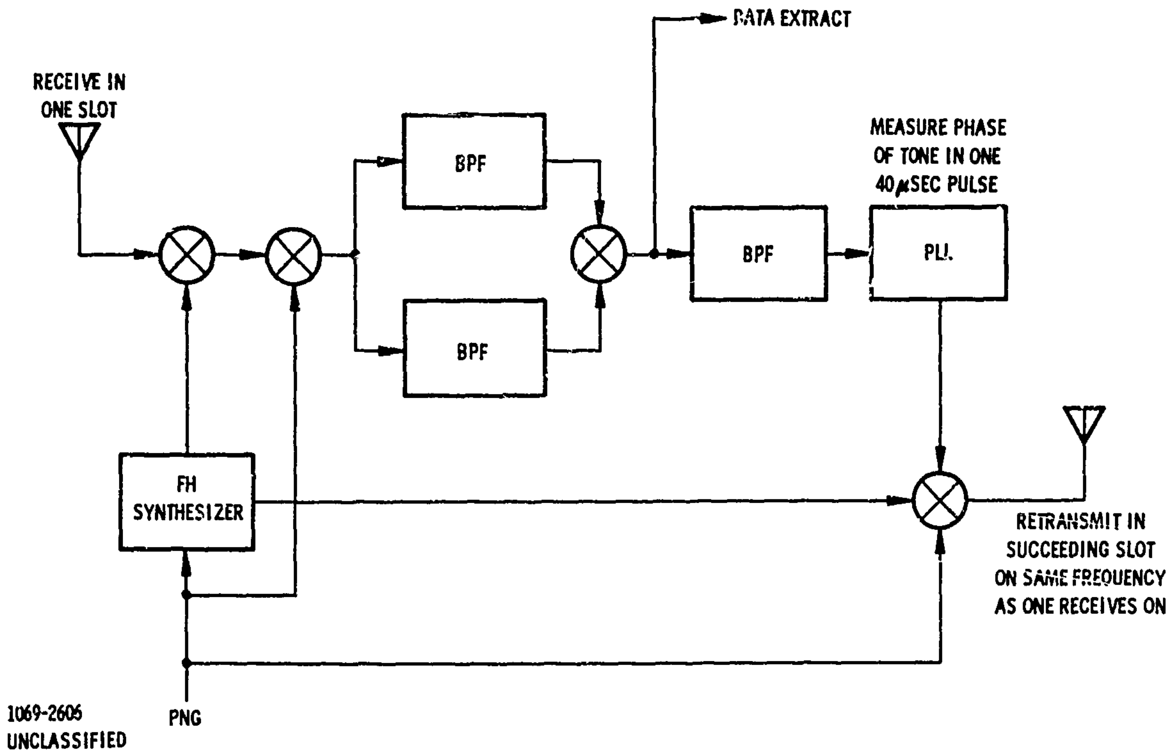

Figure 4-37. ILS Transponder Processing.

so as to reduce the jammer's power grabbing effect on the satellite. Frequency rehopping and PN recorrelation is also indicated for the retransmitted signal on the downlink primarily so as to make this signal have the same format as the rest of CNI. One also obtains processing gain in this way against interference introduced at the

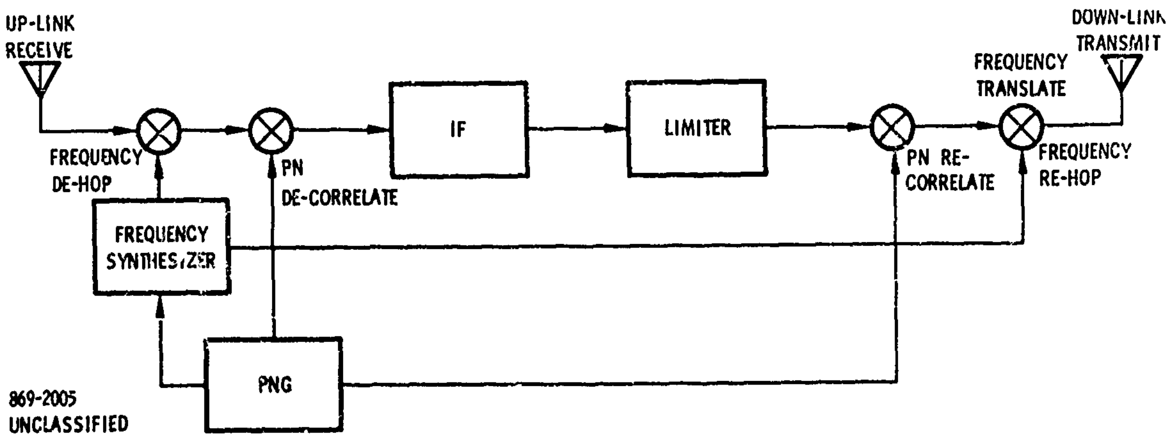

Figure 4-38. Satellite Waveform Processing (FH/PN/TH) for AJ.

receiver on the satellite downlink. A single frequency synthesizer and PN generator is indicated in Figure 4-38 for reasons of implementation simplicity in the satellite. This does, however, require that all uplink access signals arrive at the satellite in sync with the satellite PN generator to within a fraction of a PN chip. This, however, is a capability which the NAVSAT measurements allow the aircraft terminal to have so that the problem does not appear to be a limiting one.

Concerning civilian compatibility, the parameters of the modified waveform presumably provide this, based on the referenced information.* This pertains primarily to developments for collision avoidance and ILS. Now in regard to the compatibility of the waveform design in the sense that an economical civilian version can be derived from it, comments and Table 4-3 of Section 4.2.1.8 are essentially applicable to the modified waveform design except perhaps in the area of ranging. Since the design here basically does not depend on coherent sidetone ranging, but uses PN clock phase instead, then it may also be desirable to use PN for the civilian version as well. One would then retain the PN for the COM mode, realizing the protection it affords against multipath intersymbol interference and against transmitters that share the same slots.

*Diamond and Kunze, op. cit.

SECTION V

TECHNOLOGY IMPLICATIONS

The impact of hardware state of the art was investigated to determine, at least in a qualitative sense, the implications of technical feasibility for implementing the preferred waveform described above, and the resultant relative cost, complexity and time frame thus indicated. The results of the investigation could then be weighted in the final consideration of a preferred waveform and provide a technology base from which to design a "scaled down demonstration transceiver/processor" for Task A003. In order to meet the above objectives, the approach taken was to define those components critical to any waveform to be considered and investigate the state of the art of these components. Thus, this section identifies principal components (as opposed to a functional design or implementation of any waveform) for the most part and some implementation of unique techniques with general usage independent of the specific waveform chosen.

In the conduct of the investigation, four general areas were identified as critical for use in the 1-10 GHz region with application to an anticipated CNI environment:

- Signal reception
- Frequency synthesis for hopping modes
- Transmitters
- Logic components - notably LSI/MSI technology

These ultimately lead to four component categories,

- RF amplifiers
- Frequency hopping synthesizers
- High power traveling wave tubes
- Logic components

each of which are discussed in the following paragraphs.

5.1 TECHNOLOGY GAPS

Results of the investigation indicated there were gaps in four areas; namely,

- RF amplifier dynamic range
- Frequency hopping synthesizer with phase coherence
- High peak power traveling wave tube
- LSI component development

With respect to the preferred waveform, the impact of these areas initially appears to be one of development time rather than an advancement in technology. The pacing item is the high power transmitter TWT which will require approximately a one year development time. Presently a 1 KW average tube providing 100 KW of peak power can be developed either with a ring bar design traveling wave tube or a gridded crossed field device such as that developed by Warnecke Electron Tubes, Inc. Increasing peak power while maintaining a 1 KW average will rapidly result in a requirement beyond the state of the art. Physical parameters (size, weight, etc.) will increase prohibitively and thermal circuit design becomes a major factor. However, modest increases such as perhaps a 250 KW peak (1 KW average) can be accommodated providing additional circuitry such as driver development, LSI circuitry development, etc., can reduce the total system package size to a practical value. This again, is a problem of cost and time but not a technology breakthrough.

Figure 5-1 shows the trends in digital technology. As can be seen, higher speed presents a problem in power consumption and for digital rates above 1-5 MHz the cost increases significantly at present. The use of LSI to reduce system volume is done at the expense of system cost and becomes feasible only in large quantities. Therefore, small integrated systems can only be considered on the broad CNI system concept scale but not for limited quantities such as demonstration programs.

Synthesizer switching rates are comparable with designs presently being considered and the use of a phase coherent synthesizer in a demonstration unit appears within the state of the art. Although eliminated by the use of a coordinated TDMA waveform concept, the problem of amplifier dynamic range is a bona fide technology gap which is not otherwise circumvented. The following paragraphs will briefly

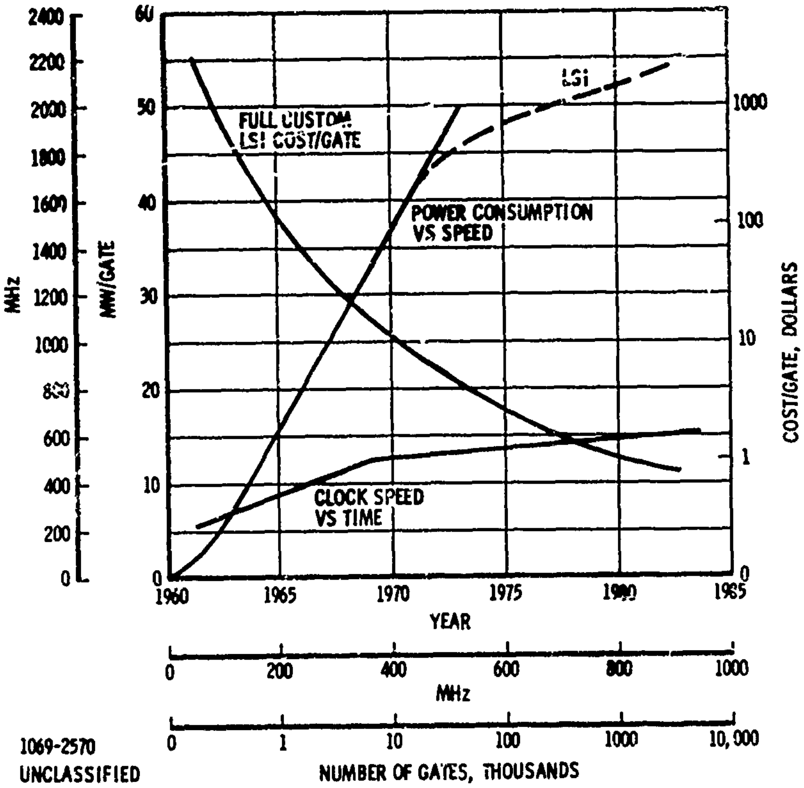

Figure 5-1. Trends in Digital Technology

discuss the four areas indicated above independent of any waveform consistent with the aforementioned objectives. More extensive discussions and further detail can be found in Section 5.2.

5.1.1 RF AMPLIFIER DYNAMIC RANGE

One of the problems associated with multiple user environments is the near-far power ratio. A system which handles multiple users simultaneously may have to process two signals with some 100 db of differential power level. Consider a frequency channelized system whereby each user occupies a discrete frequency which is pseudorandomly permuted over the total bandwidth allocated for system utilization. If an undesired signal within the bandwidth is closeby, the signal level will cause receiver saturation with the result that intermodulation products will occur which may be at a higher signal level than the desired signal. Figure 5-2 illustrates the near-far problem for 1 KW transmitters situated 100 feet (near) and 300 n miles (far) from the receiver. The maximum input levels allowed for two undesired (near) signals

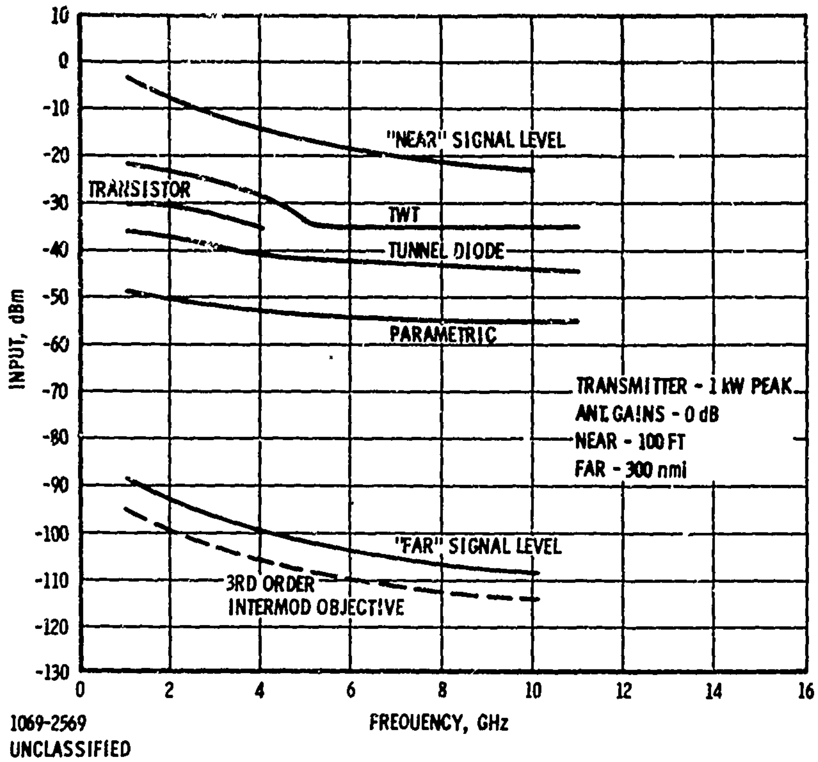

Figure 5-2. RF Amplifier Input Power for Near-Far Environment

resulting in third order intermodulation products with a signal level 5 db below the desired (far) signal is shown for the four candidate amplifiers. As can be seen, the best performance is some 15 db short resulting in saturation. The use of YIG filters for preselection is considered which can provide additional rejection outside of a 7 MHz (minimum instantaneous bandwidth) bandwidth. Another alternative would be to mix at the RF frequency (with a mixer design considered feasible by Anzac Electronic Division of Adams-Russell in Connecticut). However, the IF filter requirements to provide adjacent near-far channel rejection in this case is impractical.

5.1.2 FREQUENCY HOPPING SYNTHESIZER

High hopping rates with phase coherence is not within the state of the art. Present synthesizers with phase coherence are limited to approximately 2 millisecond switching time (for 10 MHz jump) although conceptually the switching times and hop rates could approach diode switching rates by a combination of synthesizers utilizing

the phase coherence property. This is done at the expense of size, complexity, and cost. The utility of such a scheme must be confirmed through more detailed analysis.

5.1.3 HIGH POWER TRANSMITTING TUBES

Present technology has been directed toward CW and short (≤20 μsec) pulse width designs. A requirement for a 4-6 millisecond, high power application requires a new design approach to minimize physical parameters within the performance specifications. The design utilized will affect the tube physical parameters such as size, weight, etc. A frequency in the 2.5-3.0 GHz region would more likely be optimum from a performance-size, weight point of view. Preliminary estimates indicate a tube on the order of 40 pounds in a 30-36 inch by 4 inch diameter package. No significant reduction in these parameters for the 1.5 GHz frequency range is forecast.

5.1.4 LSI COMPONENTS

The biggest impact to the CNI program with respect to the area of LSI is that of development time and money. Presently, integrated circuit manufacturers are only integrating circuit functions which are marketable on a large volume basis so as to amortize initial development costs. The result is that standard functional arrays such as shift registers, counters, etc., utilized by the computer industry are being integrated. There is a tradeoff between speed and cost with the result that MOS technology will probably be the most pursued of the possible technologies although many computer requirements are for higher speeds at present. Some functional circuits such as matched filters may eventually be forthcoming but in all probability the CNI development efforts will have to bear the costs of integrating the system designs. For extremely high data rates, the present technology is limited to standard products with high power consumption.

5.2 COMPONENT SURVEY

5.2.1 DISCUSSION OF AMPLIFIER CHARACTERISTICS FOR USE IN THE 1-10 GHz FREQUENCY BAND

Information was obtained from manufacturers of parametric amplifiers, transistor, tunnel diode and TWT amplifiers to determine their characteristics in the 1-10 GHz region. One immediate problem is to determine the specifications to which the amplifier must operate in order to determine its usefulness in the CNI waveform

implementation. Based on previous work* using the paired echo interpretation of distortion, phase linearity with ±3-5 degrees and a 100 MHz bandwidth at the 1/2 db points were chosen as the initial guidelines (see Appendix I). The characteristics of interest were the bandwidth achievable, the phase linearity and the nature of saturation. The latter refers to both the point of saturation and the characteristic of the amplifier in the saturation (1 db compression to curve elbow vicinity) region because of the near-far problem.

The near-far problem represents the situation of at least four aircraft operating in the communication network and is presented in the context of frequency channelization. Two aircraft are communicating over a large range. The aircraft receiving the signal from the far range has the other two aircraft near him operating on two additional channels independently so that, although they may be on separate channels, all three signals are within the wideband RF amplifier. If the two near aircraft are close enough to drive the amplifier into saturation, intermodulation products result which could coincide with the channel being used by the far range aircraft. The resultant intermod frequency could be greater than the "far" signal preventing communications.

For the 100 MHz bandwidth, 3-5 degree phase linearity variation requirements, all four amplifier types can be utilized. The differences become first, performance - linearity and noise figure - and secondly such things as cost, complexity, etc. The parametric amplifier provides the best noise figure performance and the TWT-type I the worst (relative). From a NF point of view a parametric amplifier would be chosen (see Figure 5-3 and Figure 5-7 for TWT) with possibly a compromise to the tunnel diode amplifier because of the temperature stability requirements, pump power supply required for the parametric amplifier and its cost. However, the concern in communication with the intermodulation distortion and the resultant possiblity of channel (adjacent) blockage due to the near/far problem results in the saturation level and linearity (gain) being of most importance. From this point of view, the TWT-Type I appears to be the best candidate. Figure 5-4 shows the input power level in dbm at which point 1 db compression is experienced.

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*DSCS Advanced Modulation System Study, Contract No. DAAB07-68-C-0346, MRL, Final Report, 15 October 1968.

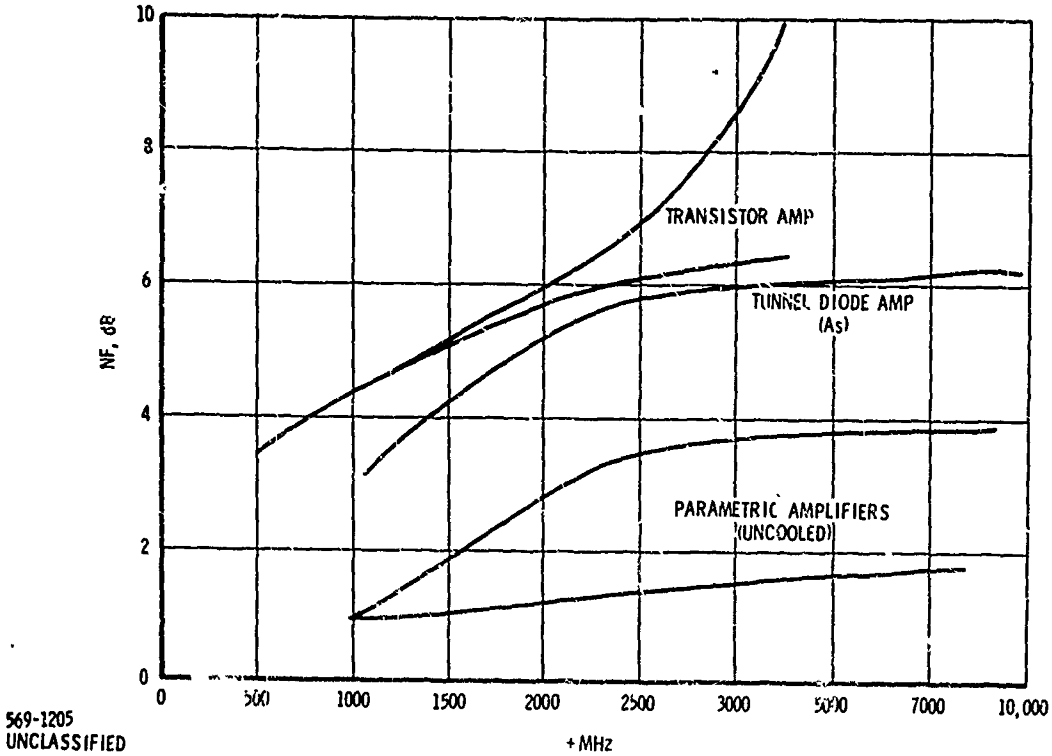

Figure 5-3. Noise Figure Range Versus Frequency

Based on Figure 5-4 the transistor amplifier offers some 10-15 db advantage over the tunnel diode amplifier and 12-17 db advantage over the parametric and Type II TWT amplifier (using the best level in each case). The Type I TWT amplifier, however, exhibits the best performance with some 5-7 db advantage over the transistor amplifier. Figure 5-4 presents some initial comparison without consideration of gain-frequency variations associated with the respective amplifiers. Perhaps, a better comparison can be made based on the intermodulation distortion levels experienced in these devices. In order to provide some basis for comparison, a common denominator needs to be established; hence, the 1 db compression point was chosen for reference. Based on information available from manufacturers concerning IM products the chart in Figure 5-5 was utilized to determine IM levels. The results were closely correlated with the information given on TWT's although the 1 db compression point had to be estimated from the general data sheets in some cases. Figure 5-5 is used by Aertech for both transistors and tunnel diodes. Many manufacturers are now identifying the IM intercept point on the component data sheet.

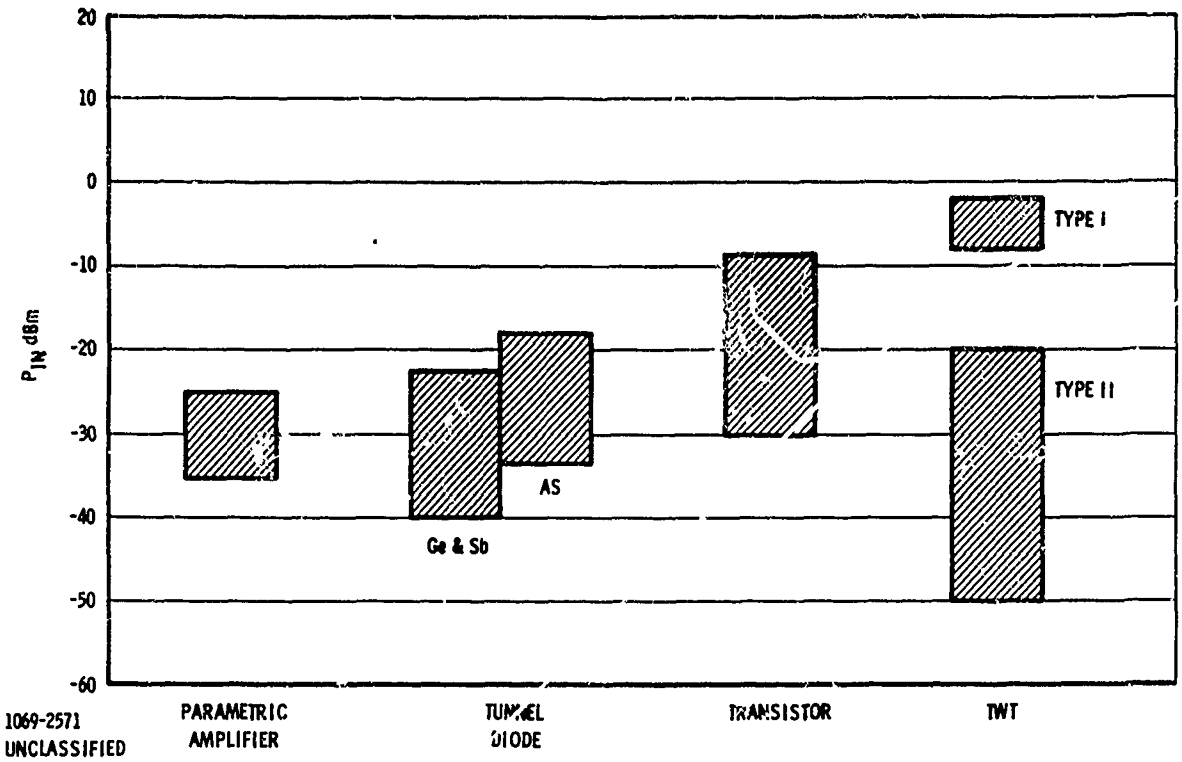

Figure 5-4. Input Power for 1 db Compression Point - Typical Value Range

Figure 5-2 was generated with the use of the chart in Figure 5-5 utilizing available data sheets and the best performing components in each case. As indicated in Figure 5-2, no amplifier has the dynamic range required to meet the environment as postulated. From an overall point of view the transistor amplifier is preferable (although the TWT cost may be comparable to the transistor) providing the restriction in frequency to the 1-3 GHz region is acceptable. The tunnel diode amplifier provides a good compromise being comparable in cost, requires little input power and models are available over the entire 1-10 GHz region of interest. The parametric amplifier offers low noise figure but at the expense of cost, high input power (depending upon pump frequency) and complexity. If system parameters can be adjusted (i.e., transmitter power, propagation margin, etc.) to within the levels for which the TWT dynamic range is sufficient, the TWT would be the only amplifier (Type I) operable within the postulated environment. It would appear to be a tradeoff between the TWT margin and the overall desirability of the transistor amplifier. From a reliability

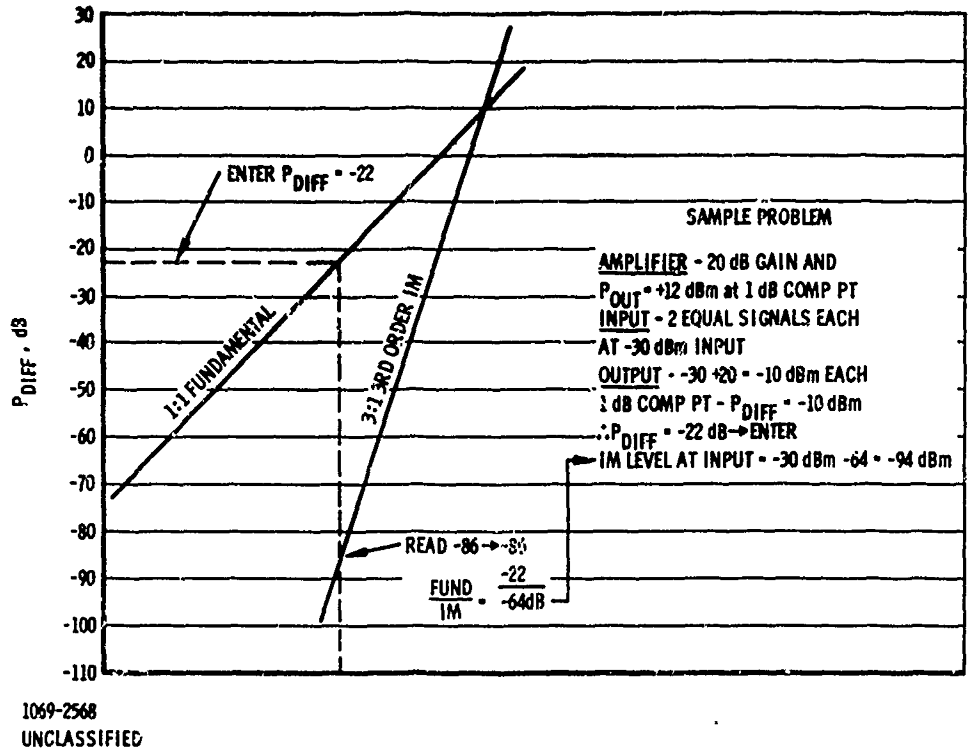

Figure 5-5. Intermodulation Chart - Aertech Industries

point of view, the transistor, tunnel diode, TWT and parametric amplifier are desirable in that order. Discussions of the aforementioned amplifiers are based on the material presented in the following paragraphs which will address the amplifiers individually.

5.2.1.1 Parametric Amplifiers

Parametric amplifiers providing a 100 MHz bandwidth (0.5 db points) are within the state of the art in the 1 - 10 GHz region. The bandwidth requirement becomes tight at 2 GHz (and below) for a one stage device and may require more than one to achieve a 0.5 db specification over 100 MHz in this frequency range should that be required. One stage typically has a 15 to 20 db amplification factor, phase linearity over the bandwidth to within ±3 to 5 degrees (10 degrees worst case), and will cost in the neighborhood of \$8000-\$10,000 for a one stage amplifier with a 50-percent increase per stage. This cost will reduce to \$2-5000 in large (100 or more) quantity. The parametric amplifier must be temperature stabilized to within

several degrees in order to maintain the phase/amplitude characteristics given above, and this is usually done by maintaining temperature at the highest ambient temperature anticipated using heater control. Present amplifiers use either a solid state device or a klystron to provide the idler frequency pump power required for operation. Solid state devices use about 5 watts of input power to operate - klystrons can require up to 100 watts. The impact in terms of paramp performance (non cooled) is that for a higher pump frequency, the noise figure is lower and at present this requires a klystron for a low NF.

Great strides have been made over the last several years in the area of low noise, broadband parametric amplifiers. The effort has been in obtaining low noise figures, packaging and solid state pumps. Thus, cryogenic cooling (closed cycle cryogenic refrigeration) and the use of high idler frequencies for use at room temperature to achieve low noise figures (electronic cooling) has been emphasized. Consider the approximate expression for excess noise temperature (T_e) of a one port parametric amplifier.

$$T_e \approx \left[\frac{f_1 f_p}{M^2} + \frac{f_1}{f_2}\right] T_D \tag{1}$$

where

f_1 = signal frequency

f_2 = idler frequency

f_p = pump frequency = $f_1 + f_2$

M = quality factor of the varactor

T_D = varactor physical temp

With respect to choice of f_2, it is found that

$$T_{e_{min}} \cong \frac{2f_1}{f_{2_{opt}}} T_D \tag{2}$$

$$\text{for } f_{2_{opt}} = M$$

In effect then, low noise is achieved by cooling cryogenically the varactor temp T_D or by raising $f_{2_{opt}}$ to obtain electronic cooling. Although f_2 is important for room temperature operation, it is of minor impact in the cryogenic case. For example, assume

$$f_1 = 4 \text{ GHz}$$
$$M = 50 \text{ GHz}$$
$$T_D = \text{(cryogenic) } 20^\circ \text{ K}$$
$$f_2 = 20 \text{ GHz}$$

Thus, from Equation (2), $T_e = 5.7^\circ$ K (amplifier only). For room temperature (290° K) T_e , 82° K. Using $f_2 = 50$ GHz, however, T_e becomes 3.6° K and 46° K for cryogenic and room temperature operation respectively. Future improvements in solid state pumps at higher frequencies is expected. The result will be smaller packages, lower power requirements and improved noise figures for paramps.

The third order intermodulation product has been measured in some para-amplifiers according to manufacturers. This is the one $(2f_1 - f_2)$ normally considered since it is the largest of the higher order products. Information from AIL indicates that the level of this product can be approximated over the range to the 1 db compression point as

$$IM_{level} = 2P_1 + P_2 + K \tag{3}$$

where

P_1 = power output of the stronger signal (db)

P_2 = power output of the stronger signal (db)

K = factor (0 to 10 db); some use 10 most use 0 db

Thus, for two signals, one at -15 dbm output and one at -75 dbm out (-95 dbm input power) the intermod would be

$$IM = 2(-15) - 75 = -105 \text{ dbm}$$

or

$$2(-15) - 75 + 10 = -95 \text{ dbm}$$

Thus, consider the case of two signals at -40 dbm whose intermod coincides with a third frequency of interest being received at -130 dbm. The IM will be

$$2(-40) - 40 + 0 = -120 \text{ or (for } K = 10) - 110 \text{ dbm}$$

Thus, the intermod will be 20 db above the wanted signal.

Parametric amplifiers thus provide operation within the specification with very low noise figures but are expensive, require temperature control and may require up to 100 watts for the pump supply. Cryogenic cooling can be utilized if low noise figures are required, however, this is impractical for satellite use and, although possible for aircraft use, present increased size, cost, and power for the refrigeration in the receiver plus the added supply for the coolant within the aircraft.

5.2.1.2 Tunnel Diode Amplifiers

Tunnel diode amplifier development has been in the area of associated circuitry-coupling networks, circulators, etc., in order to produce a reliable stable amplifier. Octave bandwidth amplifiers have been built and are available from the UHF to K band region. Noise figures vary from 3.5 db at the low end to 5-7 db in the higher region of interest here. Octave bandwidth provides a 6-10 db gain/stage with 14-17 db gain available for narrow band ($\Delta f/f_c \times 100 \approx 10\%$). However, gain linearity for wideband (octave) operation is in general 3-7 db. The 100 MHz (at the 1/2 db point) and 3-5-degree phase requirement can be met. The limiting factor in obtaining bandwidth is not the diode but the mechanism for separating the input from the amplified signal - usually a ferromagnetic circulator. The method of implementation is critical in obtaining low VSWR and load stability (see Appendix II). The type of diode utilized affects the output power level and the temperature characteristics. Three types of tunnel diodes can be used - Germanium, Gallium Arsenide (GaAs), and Gallium Antimonide (GaSb). Table 5-1 indicates some typical values.

It can be seen that the Gallium Arsenide diode presents the best performance with respect to the saturation point which is of interest in the near/far problem. Tunnel diode amplifiers present a compromise in NF with respect to the parametric amplifiers and cost approximately $2000 for a one-stage unit. They are thus lower in cost than the parametric amplifier and equal in price to the transistor amplifier

Table 5-1. Tunnel Diode Characteristics

Type of Diode	*Typical Noise Figure, db		Typical Saturation Output Power, dbm	Typical 1 db Comparison Point Input Power, dbm	
	S Band	X Band		10 db Gain	15 db Gain
GaSb	3.7	4.7	-16	-31	-40
Ge	4.7	4.7	-15	-30	-38
GaAs	5.8	7	-7	-21	-29

*Based on T_D = 10,000° K

at the 3-4 GHz region. That is because the transistor amplifier requires about 5 stages to the single TD stage at this frequency range. Tunnel diode amplifiers however must be temperature stabilized if the environment equals or exceeds 0° C and 50° C limits.

5.2.1.3 Transistor Amplifiers

Transistor amplifiers can be obtained to about the 3-4 GHz frequency range. For example, Aertech model A5802 is designed for coverage of the 2-3.5 GHz band (3 db BW) with a 10 db NF. Noise figures as low as 4 db can be obtained but range in the 2-10 db region over the UHF-4 GHz region. Figure 5-6 lists NF as a function of frequency as a general guide. Table 5-2 lists specifications for some typical transistor models with the saturation levels.

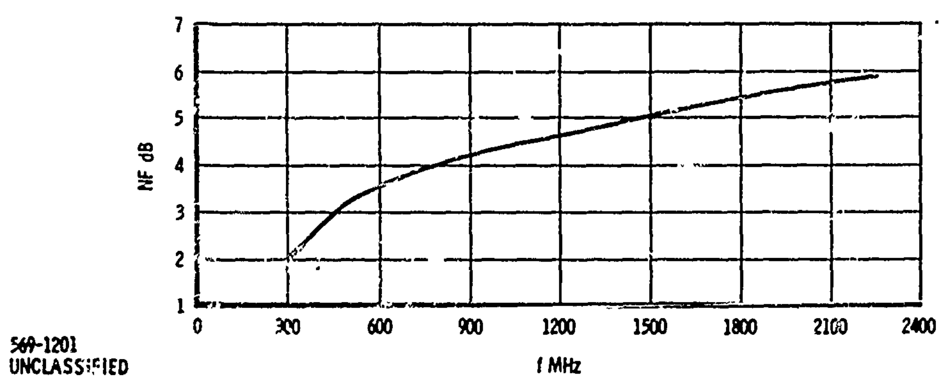

Figure 5-6. Noise Figure Versus Frequency (Trend)

Table 5-2. Transistor Specifications

Frequency, MHz	Noise Figure, db	Gain	Power Output, dbm		Phase Linearity, degrees
			1 db Comparison	Saturation	
1000-2000	6.0	15	-10	-8	±5
1435-1540	4.7	20	0	+2	5
1650-1950	5.2	20	0	+2	5
2200-2300	6.5	20	+12	+15	1

The cost of the transistor amplifier varies from $600 to $700 at UHF to approximately $2000 at the 3-4 GHz band. Phase linearity is better in the transistor amplifier than the tunnel diode or parametric amplifier, however, for a 100 MHz bandwidth, both amplifiers are comparable. The transistor amplifier provides a higher noise figure but also higher level saturation point which is desirable in the near/far environment. In addition, it is less sensitive to environment conditions and less expensive than the other two amplifiers.

5.2.1.4 Traveling Wave Amplifiers

Information obtained from manufacturers of traveling wave tube amplifiers is given here to determine the applicability of TWT's to the CNI waveform implementation. The specifications to which the amplifier must operate are the same as for the parametric amplifier, transistor, and tunnel diode amplifier discussed in paragraph 5.2.1. It has been shown that the "near-far problem" hence the level of saturation, is of major concern. Thus, this discussion will be in that context in order to facilitate comparison of relative advantages/ disadvantages of the four amplifier types.

Traveling wave amplifiers were found to exhibit two discrete areas of saturation and could be considered as two general types of tubes as depicted in Figures 5-7 and 5-8. Typically, the devices of Type II have noise figures 4 to 10 db, 15-35 db gain, and output saturation levels of -10 to +10 dbm. As can be seen, the relationship between NF and saturation level is, in general, proportional in that the higher levels exhibit higher NF. Because of the "near-far problem," the tubes of Type I are of prime interest. Although the NF is higher, the output level for saturation is 1 to 2 watts

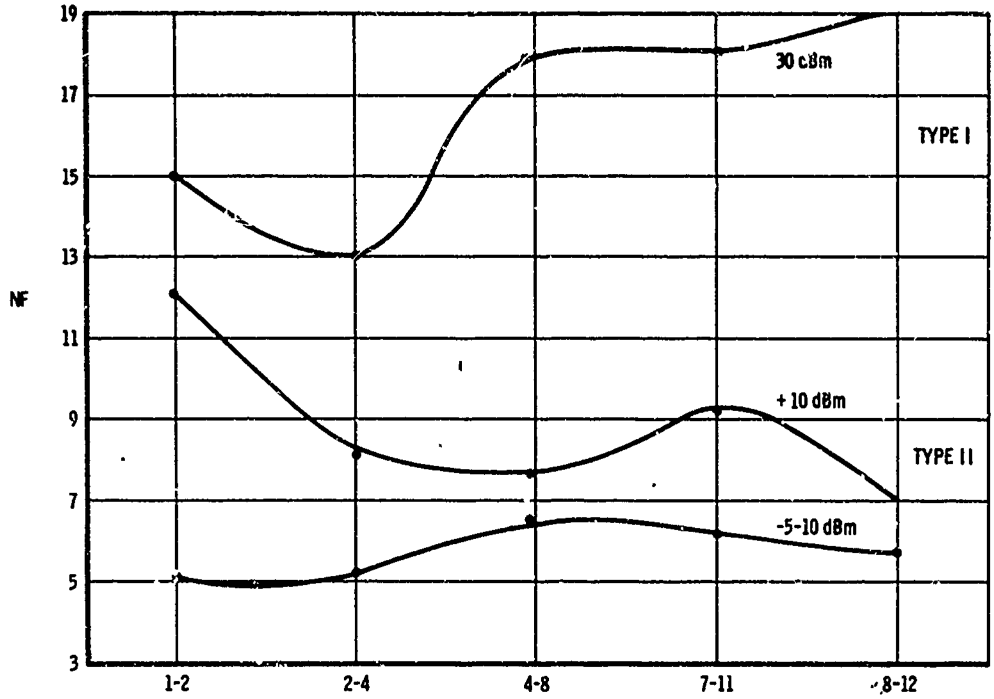

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Figure 5-7. TWT NF Versus Frequency Band (for Saturation Power Out)

which appear to be the highest level available among receiver amplifier candidates. The 10 db NF is about state of the art for these high power tubes. Figure 5-8 plots input power versus frequency rather than output because of the variance in gain and output level (saturation) available in TWT's. Despite its high level for saturation, however, the situation presented previously will still result in amplifier saturation. Assuming a 1 KW transmitted signal and a distance of 300 n miles, the power level of the received signal (0 db for antenna gain) will vary from -88 dbm at 1 GHz to -108 dbm at 10 GHz. Assume also that the level of the intermodulation products (referred to the input) should be 5 db below the received "far" signal. Figure 5-9 plots the level of input signal (i.e., the level for each of two equal input signals) for which the third order intermodulation products will be at the desired level. The desired signal is based on 1 KW at 300 nautical miles. The near signal is based on 1 KW at 100 feet and shows the relationship of that power to the required level for experiencing intermodulation products at the 5 db-below-desired signal level. Though arbitrarily chosen, the number indicates the limitation in dynamic range.

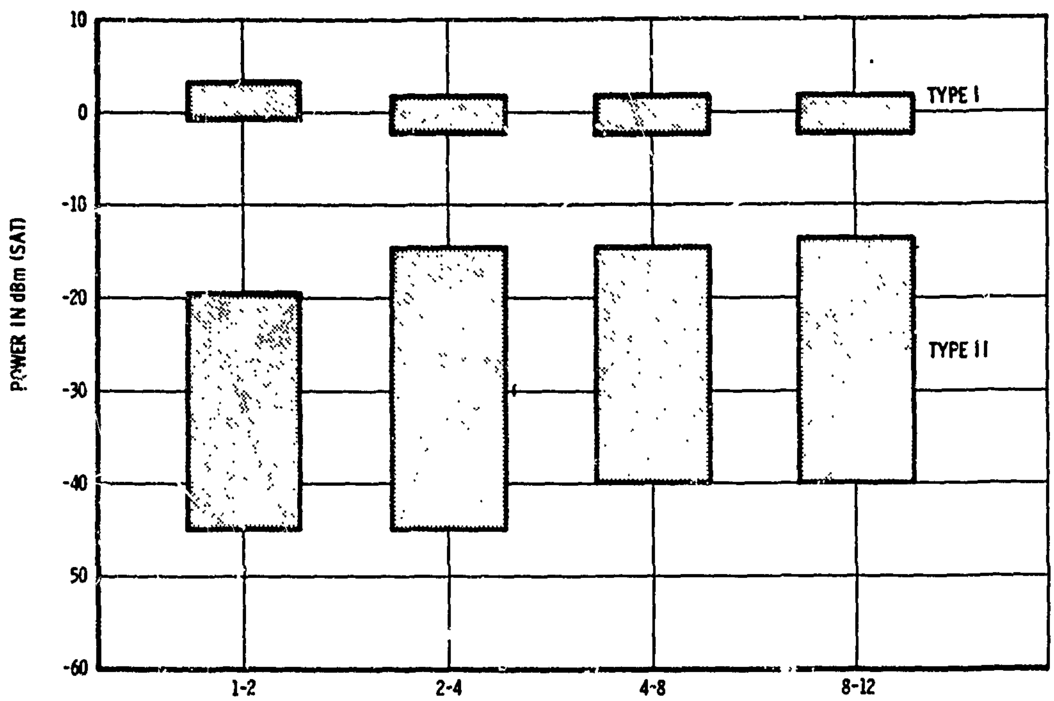

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Figure 5-8. TWT Power Input Versus Frequency Band at Saturation

Microwave Associates develops a line of devices that correspond to the Type I class of Figures 5-7 and 5-8. The NF and gain are given in Figure 5-10. Table 5-3 gives the physical dimensions.

Table 5-3. Physical Characteristics of TWT Amplifier

Tube Type	Freq. (GHz)	Length (in)	Width (in)	Height (in)	Weight (in)
2334	1-2	16.75	1.75	2.5	6.0
2330	2-4	12.0	1.5	1.62	2.5
2335	4-8	12.0	1.5	1.62	2.5
2338	7-11	11.5	1.5	1.62	2.5

All the above mentioned tubes are focused by permanent periodic magnets (PPM) and, therefore, require no power. The largest power requirement is for the

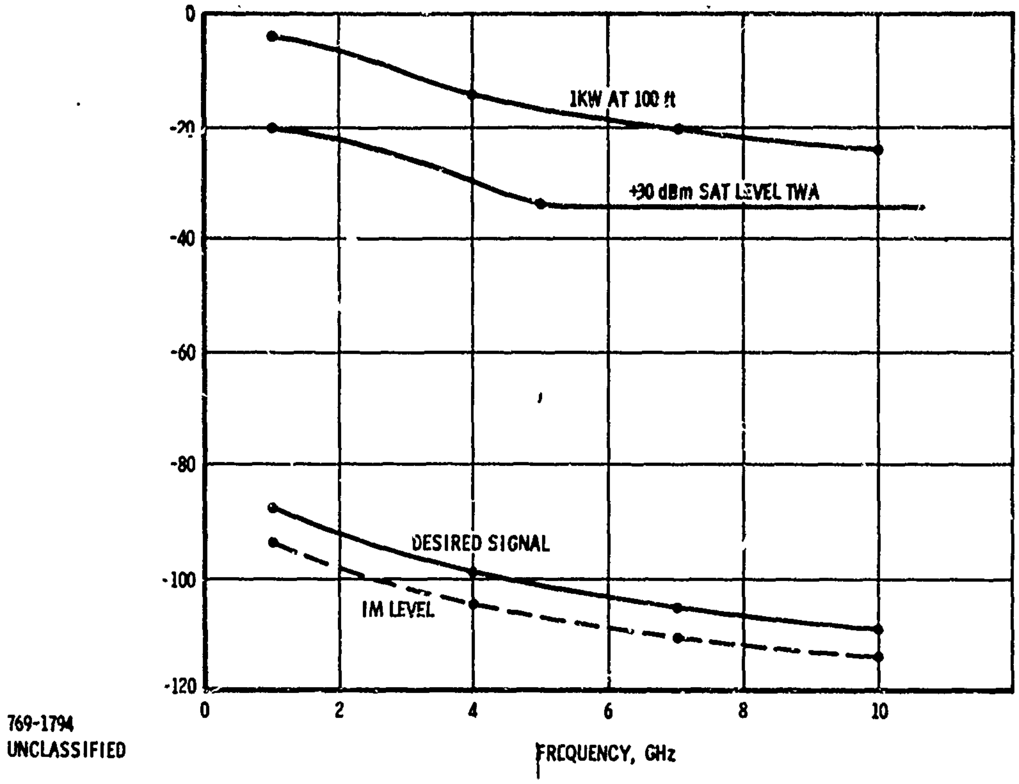

Figure 5-9. Input Power Versus Frequency for TWT to Develop IM Level

MA2334CA (Microwave Associates) which needs 700 V at 60.0 ma for the collector, 1000 V at 15 ma for the halix, 6.3 V at 1.5 A for the heater plus grid bias giving a total of 68 watts required.

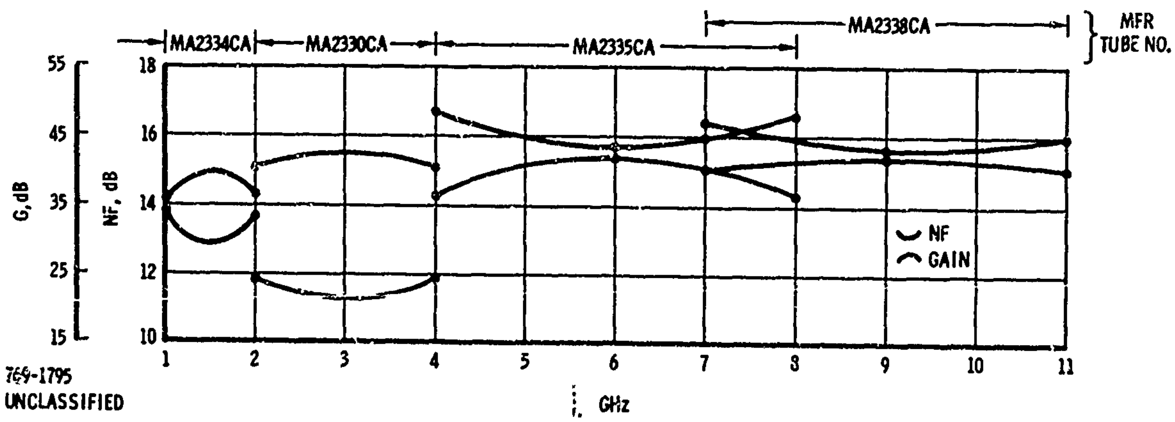

Figure 5-10. NF and Gain Versus Frequency for Tubes Listed

All the tubes are conduction cooled, may be mounted in any orientation and are typically rated for 15 g's shock and 10 g's vibration.

5.2.1.5 YIG Preselector Filters

As previously mentioned (see paragraph 5.1.1) some method of preselection could possibly provide adjacent channel rejection to bring the total resultant near/far power ratio within the dynamic range of the receiver amplifier. One of the filters considered which could be tuned over the total system bandwidth is the YIG filter (Yttrium-iron-garnet). These filters are now available (Watkins-Johnson Company, Microwave Associates, etc.) and, typically, are electrically tunable over an octave or more bandwidth in the frequency range 0.5-18 GHz.

Previous attempts at solving the problem of a nonmechanically tuned, narrow-band filter have included the use of cavities containing ferroelectrics, voltage-tuned back-biased diodes, and ferrites. In general, tuning ranges of a few percent are possible using these techniques while maintaining high unloaded Q's.

The main difficulty with these previous tuning methods is high losses, which are greatly reduced by the use of the ferromagnetic material, yttrium-iron-garnet. The unloaded Q of this material considered as a resonator is typically between 2000 and 4000 at X-band frequencies and, therefore, compares favorably with transmission-line and hollow-cavity resonators.

This approach to the problem of nonmechanical tuning makes use of the equivalence between a resonant circuit, such as inductively coupled cavity or lumped-element series resonant circuit, and an inductively coupled magnetic resonator biased with a DC magnetic field.

The tunability feature results from the fact that the resonant frequency is nearly a linear function of the DC magnetic biasing field. This feature allows a very broad tuning range, limited mainly by the bandwidth of the microwave coupling structure employed.

5.2.1.5.1 Bandwidth

The bandwidth of YIG filters can be adjusted within fairly well-defined limits. Figure 5-11 illustrates the current status of two-sphere (resonator) YIG filters. The general picture is applicable to the particular octave tuning ranges shown. Figure 5-11 also shows two areas with differing maximum bandwidths; Region B where the saturating microwave signal level will be over +10 dbm, and Region A where the input signal level for saturation may be as low as -20 dbm.

5.2.1.5.2 Insertion Loss

The insertion loss of YIG filters depend on many factors, the most important of which are the filter's bandwidth and the loss characteristic of the YIG used. Figure 5-12 displays the relationship between these factors. The type of YIG used is reflected by the limiting level and frequency range in which the filter is to be used. Accordingly, the set of curves in Figure 5-12 shows the insertion loss as a function of bandwidth with both frequency and limiting parameters.

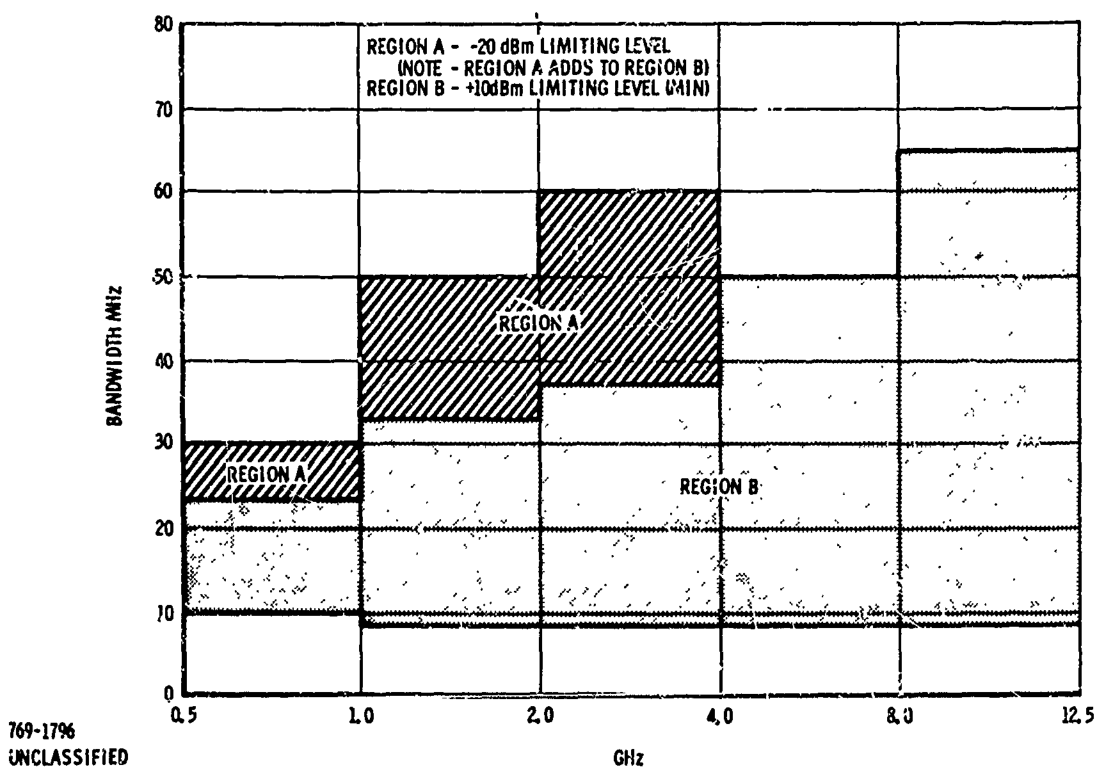

Figure 5-11. YIG Filter Bandwidth Boundaries

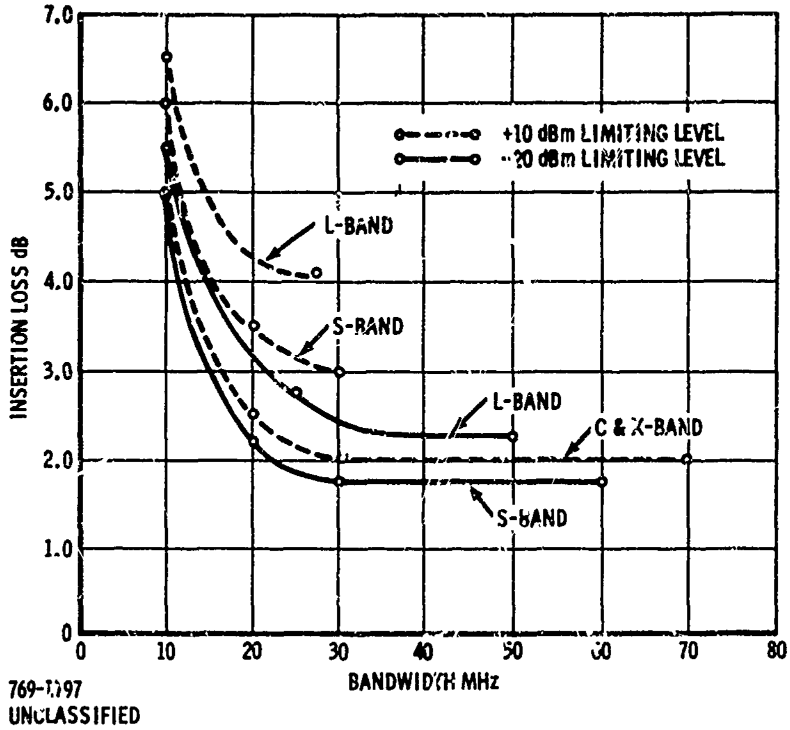

Figure 5-12. 2-Ball Coaxial YIG Filter Insertion Loss

5.2.1.5.3 Bandpass Shape and Phase-Shift Characteristics

The shape of the bandpass is controllable through the number of resonators (YIG spheres) used, and through control of the relative coupling between the elements of the filter. To a lesser degree, spurious modes and external mismatches also can influence the shape.

The skirt selectivity is determined by the number of resonators employed. For each resonator used, a 6 db/octave selectivity slope is obtained. Once the basic selectivity is established, the exact shape of the bandpass in the region between the minimum insertion loss (resonant) point and the 3 db added insertion loss points depends upon the degree of coupling between the spheres and the RF coupling structures. For the two-resonator case, the shape can be adjusted from a smooth, rounded undercoupled response to a pronounced double-peaked overcoupled response. Since the coupling between spheres and RF lines is slightly frequency-sensitive, the shape may vary over the tuning range. For example, a filter which is slightly overcoupled at the

low-frequency end of the range may change to critically-coupled at the high end of the range. For this type of bandpass shape change, Figure 5-13 illustrates the corresponding phase characteristic change. It is seen that overcoupling tends to distort the phase-characteristic at the band edges.

The depth of the off-resonance isolation is related to the number of spheres employed in the design. With 2 resonators, over 45 db is obtained; and with 4 resonators, the off-channel isolation is well over 90 db. These values are approached at the skirt selectivity rates appropriate to the number of resonators employed.

5.2.1.5.4 Tuning Characteristics

The relationship between magnetic field and resonant frequency is linear in the YIG crystal. In practical devices, the need for soft-magnetic core materials introduces two unwanted effects. First, the field at the YIG sphere is not linearly realted to the driving current if any portion of the iron either reaches saturation or has a large change in incremental permeability. Usually these non-linearities are not of importance until the frequency of interest exceeds 10 GHz. Typically, the frequency deviation from linearity for a full octave up to 8 GHz is only 2-3 MHz with no bias compensation required. In addition, below 10 GHz, special alloys which have very low retentivity are suitable which means that the hysteresis in the frequency versus tuning current curve is low, typically 1-4 MHz. In the 8-12.5 GHz region, frequency deviations due to hysteresis and non-linearity may vary as much as 8 and 40 MHz, respectively, unless elaborate compensation is employed.

5.2.1.5.5 Limiting

In the preceding paragraphs, limiting was mentioned as one of the parameters affecting the bandpass maximums. The limiting phenomena has some useful applications due to the unique properties which the filter exhibits. As previously mentioned, two levels of limiting exist which in any given filter may be frequency-related.

Using pure YIG in the 2 to 4 GHz region, the transition from low-level limiting to high-level limiting occurs in the 3.2 to 3.4 GHz region. Below 3.1 to 3.2 GHz, filter limiting occurs at input powers of about -15 dbm. When limiting occurs in the

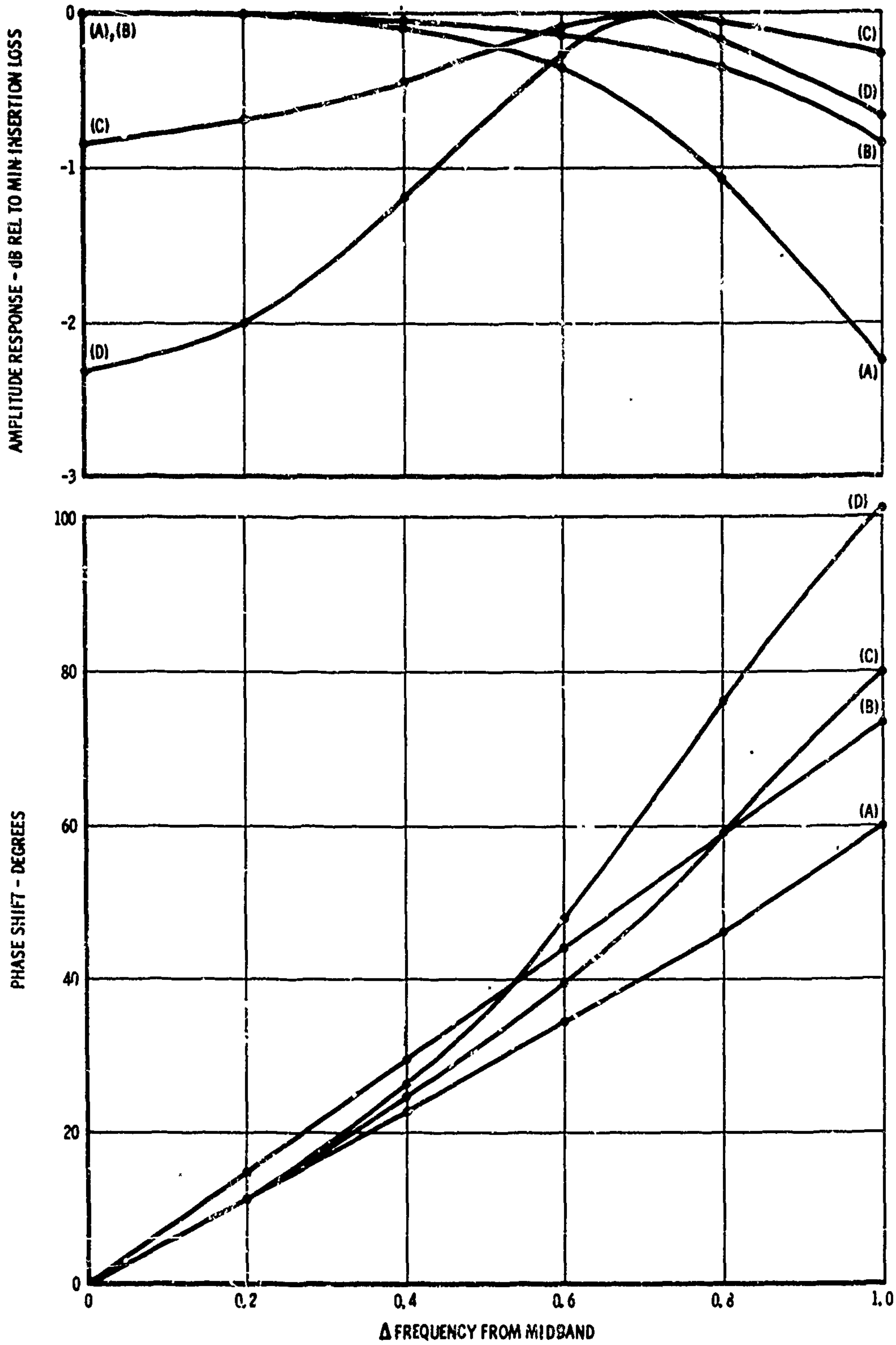

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Figure 5-13. Typical YIG Bandpass Filter Characteristics

low-level region, the limiting phenomenon is of interest since large and small signals separated only by a few MHz are selectively limited within the passband. A large signal experiencing limiting will not introduce compression of adjacent small signals, nor will it introduce mixing products provided a few megahertz separation exists between the two signals. Above 3.4 GHz, the limiting which occurs does not have this selective action and saturation of any large signal will compress low-level signals within the bandwidth.

5.2.1.5.6 Preselector Application

The YIG filter can be utilized to decrease the receiver instantaneous bandwidth. In order to maintain the amplitude and phase characteristic of $\pm$ 1/2 db/5-10 degrees from linear, respectively, the bandpass characteristic "B" in Figure 5-13 is utilized. This gives an instantaneous bandwidth capability of approximately 7-60 MHz. In order to prevent signal saturation in the YIG, the Region B (+10 dbm) is preferable which would limit the instantaneous bandwidths to approximately 7-30 MHz with 5-2 db insertion loss, respectively. Switching times in the YIG filter design can range from 50 microseconds to a millisecond and thus for slow hop systems would not appear to be a limitation.

5.2.2 FREQUENCY HOPPING SYNTHESIZERS

With the development of more sophisticated communication systems, frequency synthesizers with high switching rates and phase coherence are placing more demands on synthesizer design. Phase coherence, as discussed here, is defined as generating frequencies such that two synthesizers with identical inputs will maintain a phase difference of zero as the hopping progresses in a pseudorandom sequence. The following paragraphs discuss the present status of synthesizer development, treating phase coherent synthesizers separately. In general, the requirements on the synthesizer for the preferred waveform are within the state of the art today, however, the switching times (not average hop rate) may be marginal, requiring additional complexity in the total synthesizer design.

5.2.2.1 Frequency Hopping Synthesizer (No Phase Coherence Requirement)

A survey of available information on frequency hopping synthesizers indicates that many synthesizers have been designed for use in relatively slow hop rate systems which do not require phase coherence. There are, basically, two approaches to

frequency synthesis; namely, mix and divide or phase lock loops. A presentation of the state of the art of frequency hopping synthesizers was given to MRL in 1968 by Lincoln Labs, which with few exceptions, still represents the status of the technology today. The results of this presentation are given in paragraph 5.2.2.1.1. The few exceptions are notably the efforts by Dana Laboratories, Inc., of Irvine, California, and Barry Research, Palo Alto, California, to achieve phase coherence. There are, of course, other contributions such as the work of Saska and Hughes of Sylvania, presented at the 23rd Annual Frequency Control Symposium which, however, will not affect the conclusions to be reached concerning synthesizer technology and its impact on CNI applications.

5.2.2.1.1 Presentation by Lincoln Labs on Frequency Hopping, 25 July 1968

A presentation was given on 25 July 1968 by Ben Hutchinson of Lincoln Laboratory.*

The presentation by Hutchinson reviewed the state of the art in frequency hopping synthesizers. The first prototype of a hopper at Lincoln Lab was developed along the lines of the NRL synthesizer. This hopper provided 4,000 frequencies over 10 MHz bandwidth with a 10 microsecond switching time for application in the Lincoln Experimental Terminal. The switching time is basically faster than the commercial Hewlett-Packard synthesizer version because the design is optimized for communications applications rather than for test equipment. (In order to reduce spurious, Hewlett-Packard employs an additional mixer and filter in each stage which increases the transient time of the synthesizer.) The synthesizer is basically incoherent because of the random initial conditions in the dividers, however, Lincoln Lab has made no attempts in the direction of coherent synthesis. They have also looked at the phase-lock synthesizer and have found it useful primarily for special applications. They do not consider it to be suitable for fast switching with a large number of channels because of the inherent slowness due to the low sampling rate involved. However, it may be noted that multiple phase lock synthesizers can be conceived and are somewhat similar to the multiple mix and countdown chain in the NRL synthesizer.

*Presented for DSCS Advanced Modulation System Study, Contract No. DAAB07-68-C-0346, Magnavox Research Laboratories.

Considerable time was devoted by Hutchinson to explaining the theoretical estimation of settling time in the mix and countdown synthesizer chain. In each stage, there is envelope delay in the bandpass filter following the mixer, delay in the divider, and envelope delay through the low pass filter after the divider. The bandpass filter is the primary source of delay. As the synthesizers are built by Lincoln, the delay is cumulative from stage to stage. Considering practical designs based on avoiding spectral overlap in the mixers, it is found that a functional bandwidth in the order of 1/2 K can be achieved with reasonable envelope delay. Here K is the number of frequencies supplied, typically 4 or 8 in the Lincoln designs. It is found that the most important technique for reducing spurious is good packaging to provide proper shielding and reduce undesired stray couplings. Lincoln believes that is is possible to go to 500 MHz bandwidth with adequate spurious rejection.

In summary, the overall switching time goes up linearly with number of stages, the number of output frequencies goes up exponentially, and the bandwidth step is fixed. A multiplier can be used to expand bandwidth if spurious is sufficiently low since there is a degradation through multiplier chains.

There is some possibility in reducing switching time by using delay lines in the control leads to the various stages. This implies a fixed delay in switching but reduces dead time and thereby gets around the linear relation to number of stages mentioned above. However, this design is not demonstrated except in limited experimental work at Page.

It is considered that the upper limit for an M'ary frequency hop system is in the order of 100 kilobits/second as limited by attainable switching time in the synthesizer. This is compatible with sequential decoder implementations for M'ary systems which also have an upper speed limit of around 100 kilobits/sec.

The TATS synthesizer currently in production by Sylvania is a good example of the present state of the art. In production quantities of 100, the cost is estimated in the order of 5K to 7K. The synthesizer has a volume of 100 cubic inches, weighs 7 pounds, and requires 15 watts. It includes an internal frequency multiplier. Since

this is an existing design, new developments can certainly achieve further reductions in size and weight and power. Estimates of a factor of 2 in size and 10 in power are made, based on improved RF gating.

The summary of some existing frequency synthesizer designs is given in Table 5-4. Projected improvements on the TATS design are summarized in Table 5-5. Further improvements feasible with known techniques but still requiring development are indicated in Table 5-6. These tables present a good summary of the state of the art in frequency hopping as currently known to Lincoln Laboratory at this time.

In summary the presentation by Lincoln Laboratory indicates that the state of the art of synthesizers is quite advanced and can be considered as a practical technology available for future system applications. The most significant comment to be made about synthesizer theory is that there appear to be no surprises in the performance, and a good correlation between theoretical calculations and observed behavior in respect to switching time is attained. Spurious reduction, of course, can only be obtained by use of a pragmatic packaging technique.

5.2.2.2 Phase-Coherent Frequency Synthesizers

In general, phase-coherent frequency synthesizers may be grouped into two categories; direct and indirect frequency synthesizers. The direct synthesis technique utilizes an open loop mechanization wherein several fixed frequencies are generated from a reference oscillator and arithmetically combined to form the desired frequency. The indirect synthesis technique utilizes a feedback loop mechanization. A frequency related to the channel spacing is compared to the output of a variable frequency oscillator which has been divided until it is equal to the reference frequency. A phase locked loop causes the VCO to track the reference thereby inheriting its stability and line width. Table 5-7 summarizes the properties of these distinct types of synthesizers as they are applicable to phase-coherent frequency-hopping.

5.2.2.2.1 Direct Synthesizer

In general, due to mixing and dividing operations, the frequencies generated by the "direct" technique will not be phase-coherent in the sense defined in Section 5.2.2. To accomplish this requires compensating envelope delays of individual frequencies and assuring that identical initial conditions are maintained in divider circuits.

Table 5-4. Summary of Synthesizer (Existing) Designs

Synthesizer Parameter	LET	TATS-First Model	TATS-Production Version	Connolly Experimental Satellite Unit	Page Comm. Engineers Experimental
Stage Center Frequency, f_c	17.14 MHz	8 MHz	8 MHz	4.3 MHz	44.8 MHz
Stage Bandwidth, B	1.28 MHz	819.2 MHz	1.092 MHz	409.6 KHz	12.8 MHz
Division Ratio, K	8	4	4	4	2
Number of Stages, N	5	10	10	6	4
Component Frequency Spacing Δf	160 KHz	204.8 KHz	273.066 KHz	102.4 KHz	6.4 MHz
Maximum Switching Time	$< 10\ \mu sec$	$< 25\ \mu sec$	$< 7\ \mu sec$	$<< 1$ msec (probably $< 100\ \mu sec$)	100 Nsec
Number of Output Frequencies ($= K^N$)	$2^{15} = 32,768$	$2^{20} = 1.048,576$	$2^{20} = 1.048,576$	$2^{12} = 4096$	$2^4 = 16$
Final Frequency Multiplier Ratio, M	16	16	12	1	1
Final Output Bandwidth, (MB)	20.48 MHz	30.1072 MHz	13.1072 MHz	409.6 KHz	12.8 MHz
Minimum Final Output Frequency Increment	625 Hz	12.5 Hz	12.5 Hz	100 Hz	800 KHz
Final Output Spurious Suppression	> 40 db	> 40 db	> 60 db	> 70 db	> 45 db

Table 5-5. Improvement in TATS Design

$f_c = 32$ MHz
$B = 3.2768$ MHz ≈ 3.28 MHz
Multiplier $M = 16$, for final output bandwidth of ≈ 52 MHz
$K = 4$
$\Delta f = 819.2$ KHz
Switching Time 2 μsec maximum (with most-significant 17 bits or or less active)
$N = 11$ Stages (min output increment = 12.5 Hz, allow $\approx$ 75 bits/sec)
(This could make it to 38.4 kilobits for sure (20.30 μsec chip); probably 76.8 kilobits (10-15 μsec chip), since in the latter case only the top 12 bits (last 6 stages) would be active.
- Spurious suppression would require work, but 50 db (after multi.) is certainly within reach.
- Could go beyond this with delay-line compensation.

Table 5-6. Further-Out Synthesizer, but Quite Feasible With Known Techniques

f_c: Between 150 and 300 MHz
B: 25.6 MHz
Multiplier $M = 8$, for final output bandwidth of 200 MHz
$K = 4$
Δf: 6.4 MHz
Switching Time: Hundreds of nanoseconds, allowing data rates of hundreds of kilobits.
- Multiplier degrades spurious suppression by 18 db. 40 db suppression in final output certainly do-able, 50 db perhaps so with more work.

Table 5-7. Types of Phase-Coherent Frequency Synthesizers

Type	Advantages	Disadvantages
Direct Synthesizer	a) Frequency switching rate limited only by switch response; no settling time. b) Phase stability same as reference.	a) Elaborate filtering needed to reduce spurious and harmonic frequencies. b) Microminiaturization difficult, thereby placing volume limitation on number of channels
Indirect Synthesizer	a) No inherent filtering problems. b) Ideal for microminiaturization, as all controlled divider stages may be implemented with IC's, and the VCO, reference oscillator, and frequency and phase detection stages may be fabricated using thin film techniques.	a) For close channel spacing, low data rate or comparison frequency needed, placing severe requirements on short-term stability of VCO. b) Due to relatively narrow control loop bandwidth, longer settling times are required after channel selection, making it difficult to switch frequencies at high rates of speed.

Elaborate filtering is generally needed to reduce spurious and harmonic frequencies. This makes microminiaturization difficult; thereby, placing a volume limitation on the number of channels. The settling time during switching has been discussed above and is a function of the envelope delays in bandpass filters, and to a lesser degree, fixed delays in dividers.

5.2.2.2.2 Indirect Synthesizers

The indirect synthesizer does not have the inherent filtering problems associated with the direct synthesizer, thereby making it ideal for microminiaturization. All controlled divider stages may be implemented with IC's, and the VCO, reference oscillator, and frequency and phase detection stages may be fabricated using thin film techniques.

However, due to relatively narrow control loop bandwidth, longer settling times are required after channel selection, making it difficult to switch frequencies at high rates. Typically, these mechanizations may have a phase lock loop with high gain and can be designed for good spurious signal rejection. However, the high gain can result in phase noise which in normal applications is not troublesome.

Where the settling time becomes an appreciable part of the frequency hop rate, more elaborate means may be employed, possibly using two synthesizers, alternately switching them on after their respective switching transients have died out.

5.2.2.2.3 Synthesizer Designs

Several coherent synthesizers have been designed. Two designs are at present commercially available; one by Dana Laboratories of Irvine, California and one by Barry Research of Palo Alto, California. The latter is a modified Hewlett-Packard design which weighs approximately 200 pounds and costs $30,000. Page Communications is reported to have a coherent synthesizer also. Of the aforementioned synthesizers, little information is available on the latter two designs. The synthesizer designed and built by Dana Laboratories and a design by MRL will be described in the following paragraphs. Table 5-8 summarizes the properties of commercially available phase-coherent synthesizers.

5.2.2.2.3.1 Dana Laboratories Synthesizer - The Dana Laboratories synthesizer overcomes some of the difficulties associated with indirect digital synthesizers, since it simultaneously achieves good short term stability and can synthesize small frequency increments. Its switching time is still a function of the control loop bandwidth, and varies from less than 200 microseconds to 5 milliseconds (although specified at 5 milliseconds, approximately 2 milliseconds is the maximum observed switching time according to Dana Laboratories), being proportional to the difference between successive hopping frequencies.

The Dana Laboratories synthesizer achieves the fine resolution (1 Hz incremental spacings) by adding an additional feature to the standard "divide-by-N" phase locked synthesizer. For frequencies between N and N+1 (where N is the multiplication factor between the synthesizer stable reference frequency and the VCO frequency), the difference frequency is generated by analog means outside of the control loop and introduced into the loop to reduce the phase error to zero and lock the VCO to the desired frequency. A DC analog voltage may also be introduced into the loop corresponding to a desired initial phase condition to set any frequency to that phase condition.

Figure 5-14 is a block diagram of the synthesizer. The example treated here is that of a DC to 11 MHz synthesizer. The output is obtained by mixing the output

Table 5-8. Summary of Commercially-Available Phase-Coherent Frequency Synthesizers

Manufacturer	Equipment Number	Sweep Rate Criteria Settling Time After Frequency Switching	Hopping Bandwidth	Phase Noise
Dana Laboratories	Series 7000 Digi-phase synthesizer	For $\Delta f < 1$ MHz: <200 micro-seconds settling time. $\Delta f > 1$ MHz: 5 milliseconds maximum settling time	DC - 11 MHz	Phase Noise < 5 mrad
Barry Research	Linear Sweep Generator LSG-6	10 MHz/Second	Four 10 MHz bands from DC-60 MHz	100 milli-radians, rms maximum phase noise

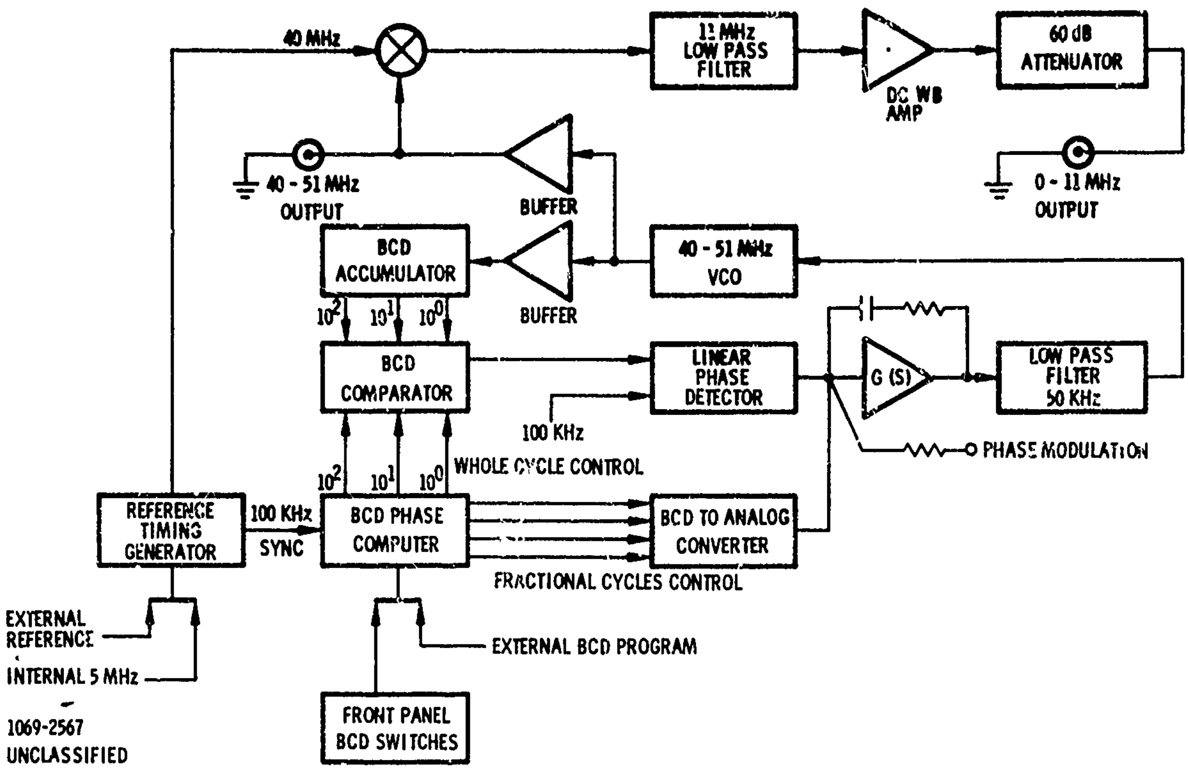

Figure 5-14. Dana Labs "Digiphase" Synthesizer

of a digital synthesizer of frequency range 40 to 51 MHz, in 1 Hz steps. The basis of the Dana Labs concept is that of computing the desired numerical value of the phase of an ideal signal (the 100 KHz reference) in units of cycles and fractional cycles, and then forcing the phase of a measured signal to exactly agree with this computed phase by means of a control loop.

To mechanize this one must note first that the measurement of integer cycles of phase can be performed digitally by simply counting axis crossings of an assumed sinusoidal incoming waveform. The interval remaining between the last axis crossing and a convenient fixed frequency clock yields a time delay. This delay can be converted into an analog quantity proportional to the fractional cycle remainder of the measured phase. By being able to digitally measure integer cycles and analogly measure fractional cycles, and being able to numerically compute their ideal values on a rapid basis, one can compare the two values and obtain a sampled data error signal to be supplied to a correction loop.

Since it is generally not desired to simply have a fixed phase output, but one of a constant phase-rate or frequency, the phase computer portion must digitally integrate the programmed frequency over a sufficiently long interval, and store the result of this computation in a phase control register. This computation must be updated at a rate equal to the sample rate in the control loop. In the example described here the rate is 100 KHz. The front panel frequency is a seven digit decimal number with 1 Hz resolution, and when integrated the numerical value of the phase has a minimum increment of 10^{-5} cycles, for the 10 μsec clock interval. Computed phase value is increased or incremented from between 400 and 510 integer cycles every 10 μsec to span the range of 40 to 51 MHz. The remaining fractional cycle portion of the computed phase contains a numerical value which should correspond to the analog output obtained from the phase detector if the input frequency and phase are correct as generated by the loop.

5.2.2.2.3.2 Conceptual Design of Phase Coherent Frequency Synthesizer - The MRL synthesizer design utilizes the properties of both the phase lock loop and direct synthesis to obtain a spectrum of frequencies which are phase coherent. Figure 5-15 shows the general approach.

The PFS (Programmable Frequency Synthesizer) is utilized to select the desired signal from the spectrum available in the spectrum generator - hence the implementation appears as a high pass filter to the PFS and a low pass to the spectrum generator. The successive combination of the respective outputs eliminates the noise associated with the PFS (preserving the selectivity) while maintaining the purity of the spectrum generator. In essence the role of the PFS could be visualized as a tunable low pass filter operating on the spectrum generator output.

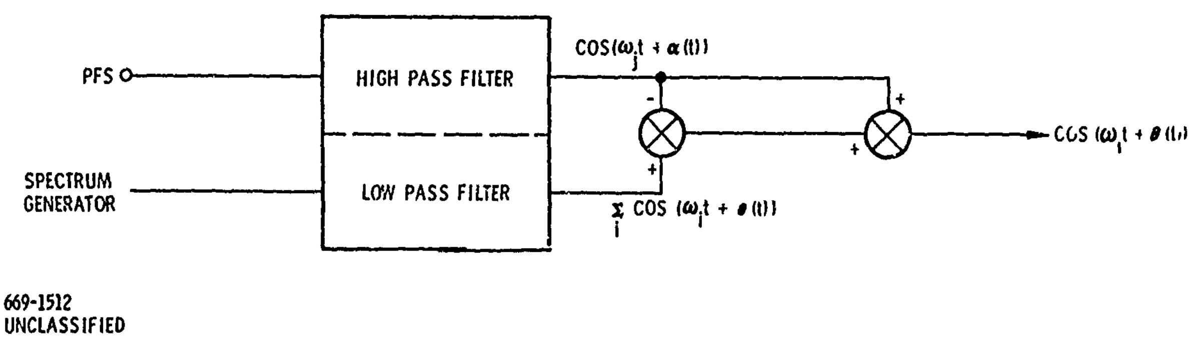

Figure 5-15. Conceptual Approach

The PFS (see Figure 5-16) is basically a phase lock loop which synchronizes the VCO output to the reference frequency. In this case the reference is set to provide the proper spacing of frequencies which are selectable by controlling the VCO output. This selection is done by changing the value of the mod-N counter since the VCO output frequency is some multiple of the reference frequency (i.e., N times the reference frequency). Thus, through controlling logic circuitry the interconnection of the mod-N counter different frequencies are generated. This control logic can be controlled by a pseudonoise coder to provide programmable frequencies on a random basis. The desirable feature of the PFS is its ability to provide a spectrum of frequencies - one at a time.

The spectrum generator provides a spectrum of frequencies over a selected band with a spacing specified by the output of the "divide by M" circuit (see Figure 5-17). The value of the multiplier (K) selects this band. The resultant output is a pulsed burst of energy at the frequency selected by the value of K with a repetition frequency set by the value of M. For example, suppose that K is set to provide a 60 MHz output and M is set for 200 KHz. Then the spectrum would be a pulsed burst about 60 MHz every 200 KHz providing a spectrum of frequencies. (For a 5 MHz reference, 55, 60, and 65 MHz and frequencies in between spaced 200 KHz would be present which could then be band limited to a 55-65 MHz region for example.) The desirable feature of the generator is the frequency purity, however, it consists of a set of spectral lines whereas only one is required at any one time.

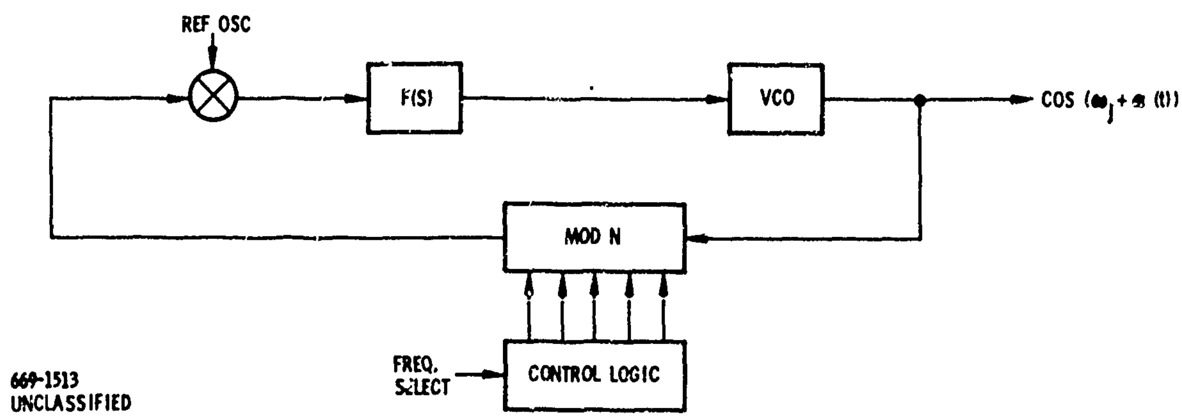

Figure 5-16. Programmable Frequency Synthesizer

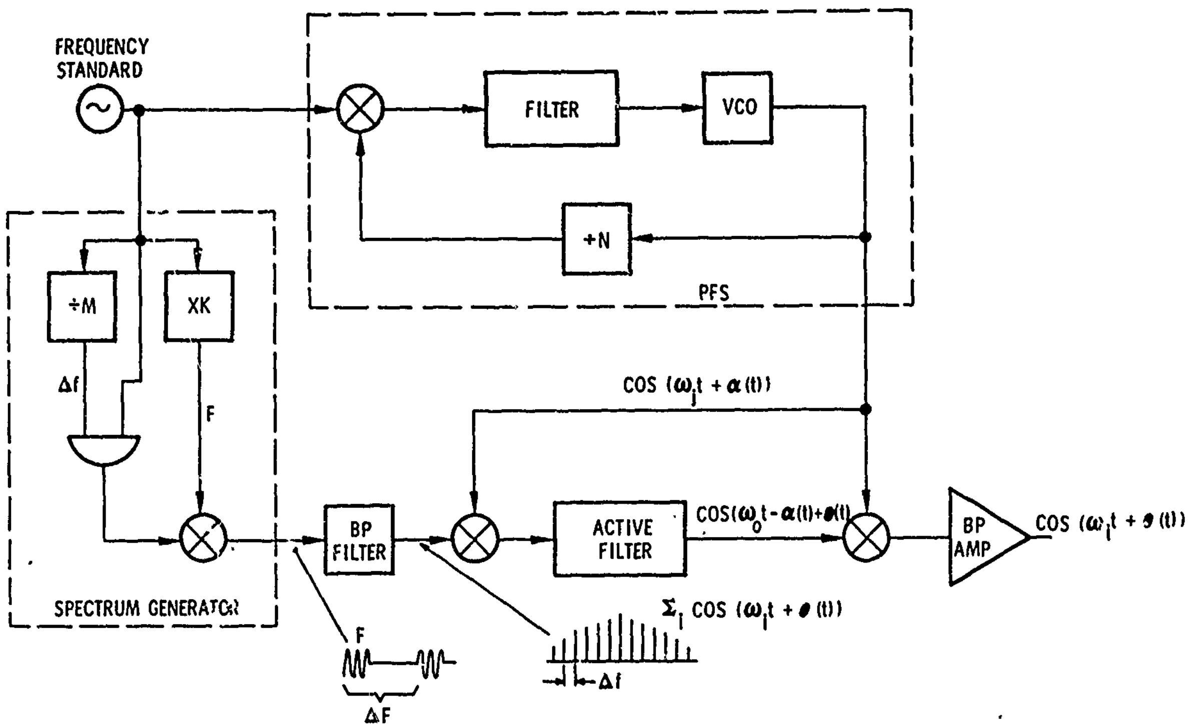

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Figure 5-17. Synthesizer Mechanization Block Diagram

By combining the spectrum generator and PFS as shown in Figure 5-17, a phase coherent output can be obtained which theoretically has tolerable phase noise yet is programmable over some range of frequencies. This can be shown as follows. The output of the spectrum generator can be represented as

$$f(SG) = \sum_i \cos(\omega_i t + \theta(t)) \tag{4}$$

$\omega_i + \omega_{i+1}$ = Δf (channel spacing)

$\theta(t)$ = phase noise inherent in the frequency standard and spectrum generator circuitry

The output of the PFS can be represented by

$$f(PFS) = \sum_j \cos(\omega_j t + \alpha(t)) \tag{5}$$

$$\omega_j - \omega_{j+1} = \Delta f$$

$$\alpha(t) = \Delta\phi \cos \omega_{\alpha(t)} t$$

where

$|\Delta\phi| > \xi$ - system phase noise specification

and

$(|\theta(t)| < \xi)$

The active filter is set to an intermediate frequency ω_o and a bandwidth BW where

$$\omega_o = \omega_i - \omega_j \tag{6}$$

$$f_{\alpha(t)} < BW < \Delta f$$

Thus, by selecting some ω_j a frequency ω_i is specified from all those available. For example, assume that the value of ω_i varies over the 55-65 MHz band and the PFS varies over 35-45 MHz (all spaced every 200 KHz). If ω_j is 40 MHz the 60 MHz line is selected which passes through the filter (ω_o = 20 MHz) with a 2 KHz bandwidth. The 2 KHz bandwidth rejects adjacent channels. In order to eliminate $\alpha(t)$ consider the filter output.

$$f_{out}(\text{filter}) = \cos \omega_o t + \theta(t) - \alpha(t)$$

Since $\omega_{\alpha(t)}$ is typically in the audio range the 2 KHz bandwidth will pass $\alpha(t)$. This output is then combined with f(PFS) which is then passed through a band pass amplifier which is set to accept the sum signal (in the example the band pass is 55-65 MHz). Then the synthesizer output is

$$e_o = \cos \omega_i t + \theta(t) \quad \omega_i = f(\omega_i) = f(\text{mod } N)$$

which is the desired result. This result is valid if the filter time delay is small compared to $\omega_{\alpha(t)}$. When this is not true, some degradation of $\theta(t)$ can result. In order to minimize distortion the value of Δf and $\omega_{\alpha(t)}$ must be chosen to ensure $T = \Delta f/BW_{filter} >> 1$.

5.2.3 HIGH PEAK POWER TWT TRANSMITTER TUBE

A survey of TWT manufacturers was made to determine the possibility of a high peak power tube with a pulse duration of 4 to 6 milliseconds. Current technology has been directed toward radar and communications applications resulting in design development at both ends of the pulse duration spectrum; i.e., CW and short pulse (up to 20 microseconds - in some cases 500 microseconds). One of the tradeoffs in the consideration of TDMA waveform selection is the possibility of extending the time slot duration from a 60 microsecond interval to intervals on the order of 4 to 6 milliseconds. Because of the random selection of slots within a frame, two slots could occur back to back resulting in a design requirement for the tube of 4-6 milliseconds with a worst case of 8-12 millisecond pulse duration. As the pulse duration increases from several μsec to milliseconds, thermal heating in the slow wave structure becomes significant resulting in a need for a more massive tube design to dissipate the heat. The crossover point beyond which thermal heating builds up to a significant amount occurs in the 100 to 500 μsec pulse duration region. In addition, these longer pulses tend to cause hot spots on the helix with periodic permanent magnet design (for beam focusing) resulting in the use of solenoids which further increase the weight and power supply requirements. Thus a tradeoff between the tube's physical parameters and the pulse duration requirement can be made to optimize the design for airborne and space applications. Although power (and pulse duration) can be traded off with weight and gain with size (length), there are bounds due to the frequency of operation, basic structure and focusing requirements, etc. Preliminary estimates by industry appear to size out the minimum tube parameters as one of 40 pounds in weight, 36 inches in length with a 30-50 db gain. Such a tube will require engineering development but is within the state of the art.

Future improvements in tube construction, focusing methods, etc. should decrease the weight and increase tube efficiency. New developments such as gridded cross field devices (GCFA) recently developed by Warnecke Electron Tubes, Inc., may

provide a better packaging form factor but present GCFA gains require a TWT driver such that total pack parameters may not significantly change. A traveling wave tube today with approximately a one year engineering development would have the following specifications:

Frequency of Interest	1.5 GHz (1-2 GHz)
Peak Power	100 Kw
Duty Factor	1% or less
Pulse Duration	to 12 milliseconds
Efficiency	30% to 40% (with depressed collector)
Gain	30 to 50 db
Weight	nominally 40 pounds
Length	36 inches (50 db gain)
Diameter	4 inches with cooling (3 inches with magnets)
Beam Focus	periodic permanent magnets

5.2.3.1 Tube Design

Tubes can vary in size from 30+ inches to six feet and in weight from 35 to 200 pounds, depending upon gain and wave structure design. Coupled cavity designs require more weight and size than a ring bar slow wave helix design. With the use of the latter design, weights on the order of 35-65 pounds can be achieved depending upon beam focusing requirements. Care must be taken in the design to form the beam down the structure and prevent hot spots from occurring on the helix. This may require a solenoid, however, the consensus is that periodic permanent magnets can be utilized for the 100 Kw peak power. This would have to be confirmed in the design effort. Tube size is proportional to gain, therefore, an increase from 30 db to 50 db would result in an increase in axial length and some weight for the focusing magnets. A tradeoff would be made between high gain (50 db) with the use of a solid state driver and lower gain with the use of a higher power driver to obtain the optimum package in size, weight, and overall efficiency. Depressed collector design (operating the collector near the cathode potential) results in an increase of some 7% to 10% in efficiency over the basic 30% efficiency of the tube, however, a TWT driver could be expected to have a somewhat lower efficiency, hence the solid state driver would be desirable. Although the tube size, weight and efficiency will be minimized in a design tradeoff, the frequency

of 1.5 GHz and the high peak power will limit the tube to approximately the 35 to 40 pounds, 30 inch length at the present time. It is felt that from the tube design point of view, 1.5 GHz is not the optimum frequency for minimum physical dimensions. Approximately 2.5 to 3.0 GHz would be the optimum frequency which would result in a reduction of 25% in weight and size over the 1.5 GHz design for the same set of requirements. The exact frequency and reduction factor of course could be determined in a design effort as one of the tradeoffs is desired.

The pulse duration requirement will require a sizeable cathode capable of such emissions over the 4 to 12 millisecond duration postulated. Thus a trade in peak power versus average power results in significant increases in the tube's physical parameters. Table 5-9 lists some of the characteristics of a 1 Kw, CW tube with those postulated for a 100 Kw peak, 1% duty factor tube.

Table 5-9. TWT Tube Parameters

	1 KW Tube	High Power Tube
Frequency Range	1-2 GHz	1-2 GHz
Duty Cycle	100%	1%
Saturated Power Out	1800 watts	100 Kw
Saturated Gain	30 db	50 db
Collector/Helix Voltages	~3.6 Kv	~50 Kv
Efficiency	15%	37%
Weight	15 pounds	45 pounds
Length	20 inches	36 inches
Total Outside Diameter	2.5 inches	4 inches

5.2.4 LOGIC COMPONENTS - LSI/MSI

Present availability of logic components are primarily of the MOS type for large size shift registers and memories (50 bits or more) because of its ability to provide a number of bits efficiently in terms of size and power dissipation. Although considerable progress is being made along the lines of "circuit function" integration; i.e., from standard (SSI) scale to medium (MSI) or large (LSI) scale integration, it is

done on the basis of general industrial usage. The development of LSI packages requires a high initial cost which must be amortized over a large quantity thus implying either a large amount of user equipment or a highly repetitive type circuit. The computer and computer related industries are, by far, the largest users of integrated circuits; therefore, the emphasis for integration is on counters and shift registers. Figure 5-18 illustrates the criteria in choosing the type of logic circuitry. The figure represents the percentage of standard, MSI, LSI, etc. utilized in a particular system with respect to number of gates. This is based on cost, reliability maintenance and particular constraints that may require a particular choice and presents a trend rather than specific usage per se. The transition from the standard product can cover limited integration (hybrid), available arrays with flexibility in performance due to various interconnections (which can be requested), to full scale integration of circuit functions. The impact on system cost is shown in Figure 5-19 which relates three possible approaches in terms of cost per gate versus number of gates. Again high usage is implied for economic utilization.

5.2.4.1 Trends

As stated previously, the large memory/shift register approaches use MOS technology which inherently limits the clock speed capability to approximately 5 MHz. Presently available shift registers now have some 256 bits. Figure 5-20 shows the projected trend of register sizes over the years. With increased integration, the power consumption will decrease and larger size registers are forecast. High speed

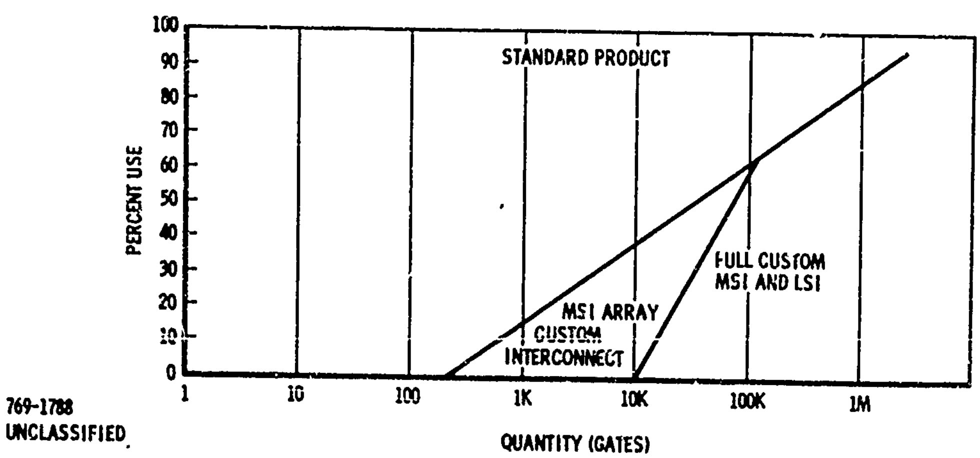

Figure 5-18. Integrated Circuit Approach Versus Quantity

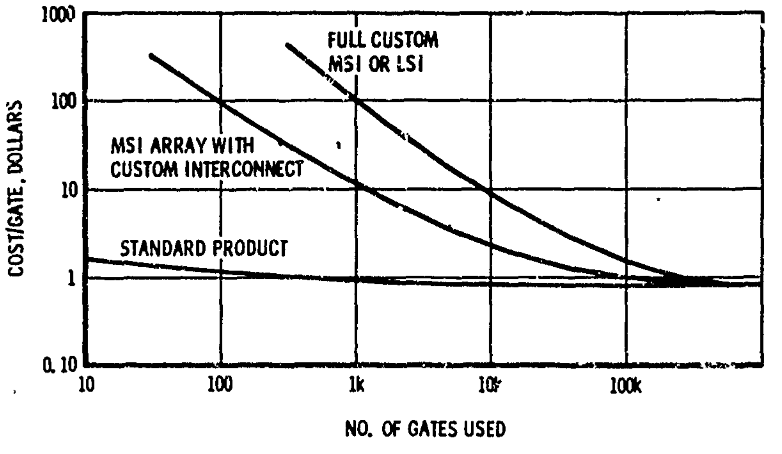

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Figure 5-19. Cost Per Gate Versus Number of Gates

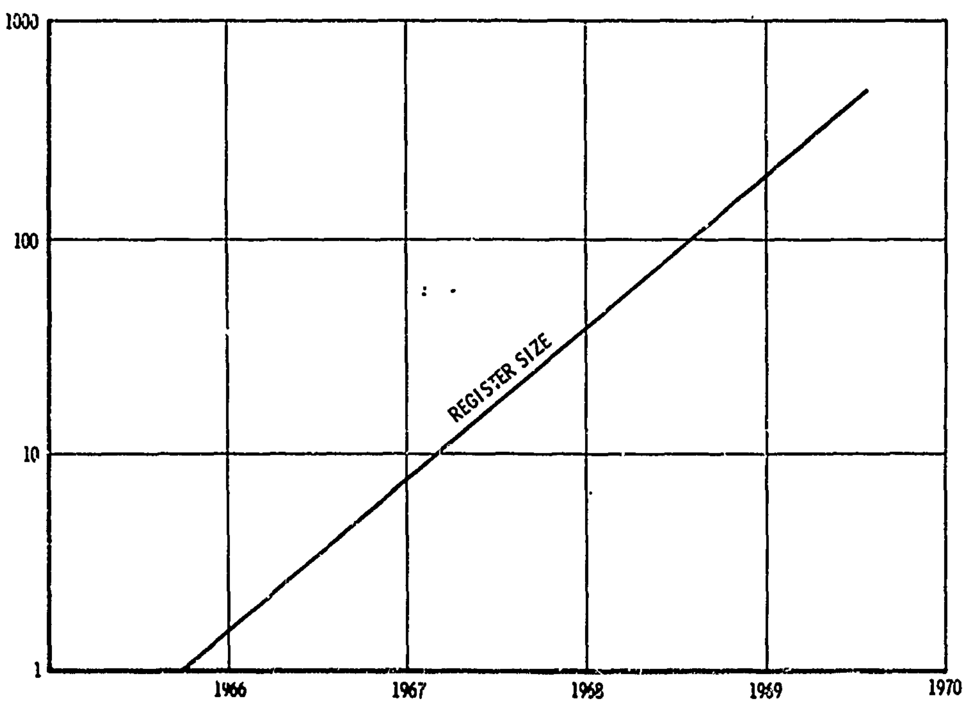

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Figure 5-20. Register Size Versus Time

logic of the MECL III Type has reached 500 MHz in theory, with shift register reliably clocking at 250-300 MHz. However, as shown in Figure 5-21, the power dissipation increases with speed. It is anticipated that, although careful integration can again minimize total power dissipation requirements by using lower level internal gates, speeds will not be greatly increased in the next few years. This is because of the need for new techniques in interconnection, such as transmission lines, etc., for high frequencies, and associated problems which will probably slow down, momentarily at any rate, the slope of increased speed versus time.

Thus, in summary, standard arrays such as shift registers, counter, standard logic gates array with custom interconnection will be available in increasing capacity for speed and bit length but specialized circuits, such as digital phase lock loops as a case in point for communications, will not be readily developed by IC manufacturers in LSI form.

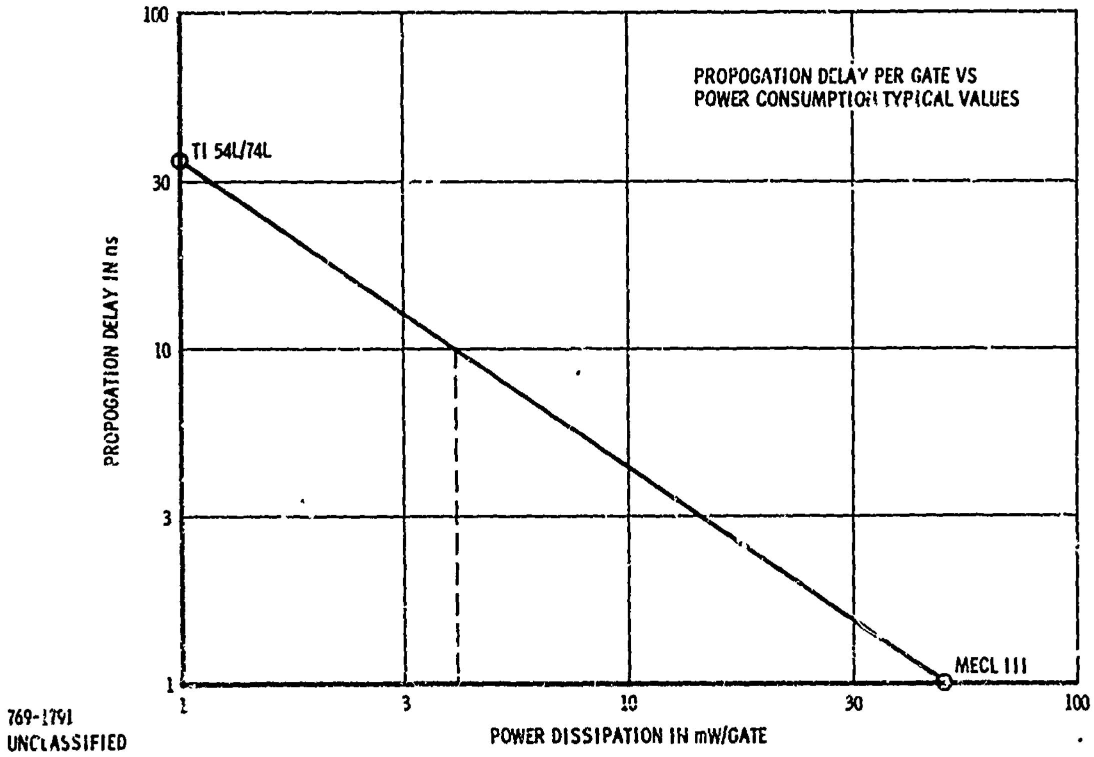

Figure 5-21. Propagation Delay Per Gate Versus Power Consumption Typical Values

5.2.4.2 Impact on System Design

Several unique implementations were investigated to determine the degree of difficulty in the mechanization and the resultant impact on integrated circuit design. The following circuits are typical of the non-standard (in today's IC world) circuit functions that could be integrated on an LSI basis to provide small size, weight and cost for the total ICNI system.

5.2.4.2.1 Digital Matched Filter

An integrated circuit implementation of a digital matched filter has been made practical through the realization of large scale integration (LSI). Figure 5-22 shows a practical integratable basic element of a digital matched filter. The element contains two registers, one for the quantized input signal and one for the replica signal, and in addition, a series of modulo-two adders and a summer. These elements can be connected in series to obtain the desired filter length and in parallel to accommodate multilevel quantization. For the case of multilevel quantization, the sum outputs are weighted depending on the quantized value of the bits input to the register.

To achieve the most bits per integrated match filter element, the logical choice of semiconductor technology is MOS. Presently available MOS integrated circuits provide dual 100 bit shift registers on a single chip. The complexity of a modulo-two adder is approximately equal to a shift register element, likewise the complexity of the summer is about the same as the complexity of n shift register elements, where n is the number of inputs to the summer. Thus, a single integrated circuit chip that accomplishes the 50 bit matched filter function shown in Figure 5-22 should be about the

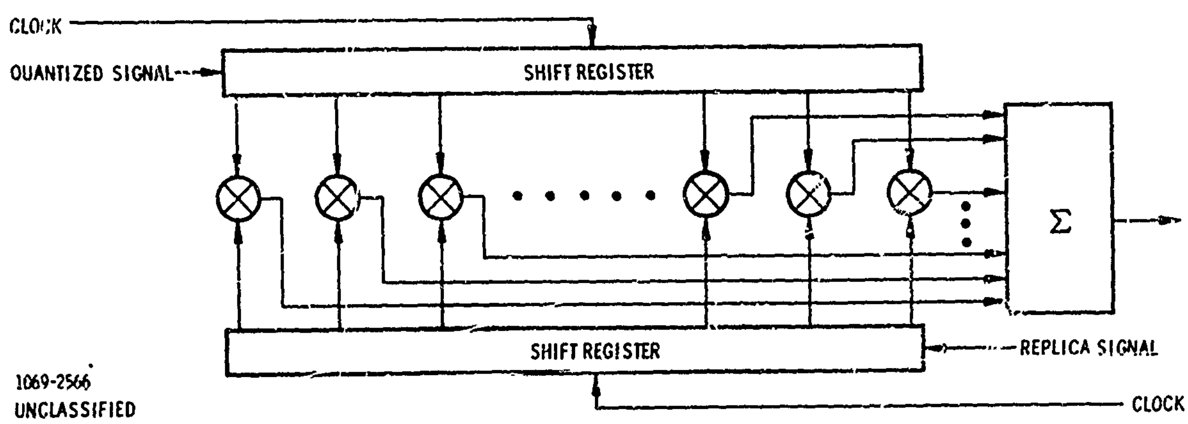

Figure 5-22. Matched Filter Element

same in complexity as a 200 bit shift register. The price paid for this high density is operating speed. The majority of the dual 100 bit registers have a maximum shift rate of 2 megabits/second, although one manufacturer claims 100 megabits/second for a dual 50 bit register. Therefore, optimistically, one might expect to obtain a 10 megabit per second matched filter element that is 50 bits in length. For higher frequency operation, bipolar semiconductors must be employed and from complex circuits available one might expect 16 bits of match filter element on a single chip with shifting rates of 30 megabits/second.

5.2.4.2.2 Digital Phase Lock Loop

The following summarizes an approach* to the mechanization of a digital phase lock loop (PLL) is shown along with the design equations; in addition, the PLL is extended to a Costas loop.

For purposes of explanation, a first order PLL is shown in Figure 5-23 and the waveforms associated with this loop as shown in Figure 5-24. This digital PLL implementation requires that the input signal be hardlimited and enter the loop as a square wave. The input square wave is mod-2 added to the output signal. The output signal matches the input signal in frequency but not in phase. Because of the phase offset a control signal appears at the output of the mod 2 adder. The duty cycle of the control signal determines the number of pulses that drive the counter from frequency sources g_1 and f_1. When the control signal is in the true state clock pulses from generator f_1 drive the counter and when the control is in the false state clock pulses from generator g_1 drive the counter. The counted mix of pulses from generator g_1 and f_1 match the frequency of the input signal.

A second order loop is shown in Figure 5-25. The second order loop is accomplished by adding to the first order loop another counter and frequency source. As before, f_1 and g_1 drive counter one according to the duty cycle of the control signal but when a count equal to M is reached in counter two, the contents of counter one are transferred to counter two, counter one is reset, and the state of the output flip-flop is changed. When the loop is locked the count of M is reached in counter two

* G. Pasternack and R. L. Whalin, "Analysis and Synthesis of a Digital Phase Lock Loop for FM Demodulation," Bell System Technical Journal, December 1968, pp. 2207-2237.

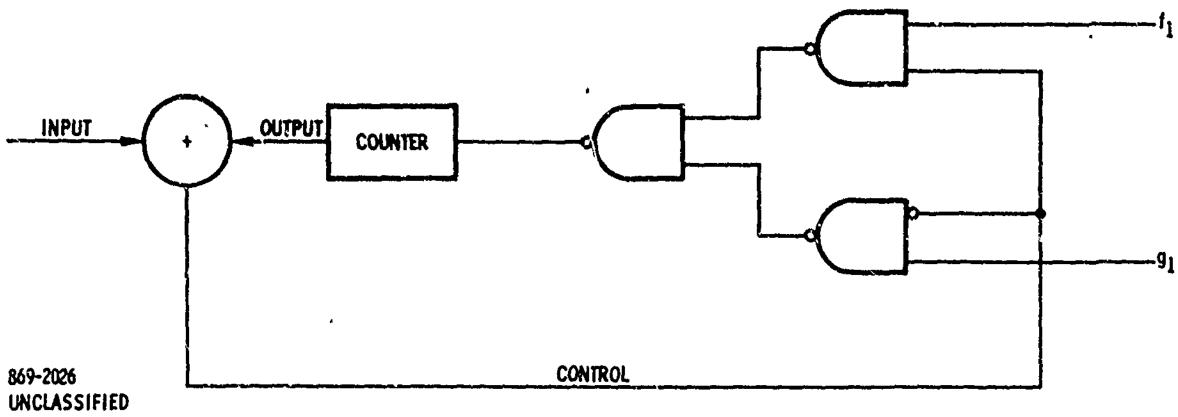

Figure 5-23. First Order Digital Phase Lock Loop

every half period. Counter two is driven by frequency f_2 when the control signal is in the true state.

In the reference, the equations relating the digital PLL parameters to the analog PLL are developed for an analog loop with characteristic equation

$$s^2 + \sqrt{2}\,\omega_n + \omega_n^2 = 0$$

It is shown that the equations for f_1, g_1, and f_2 are:

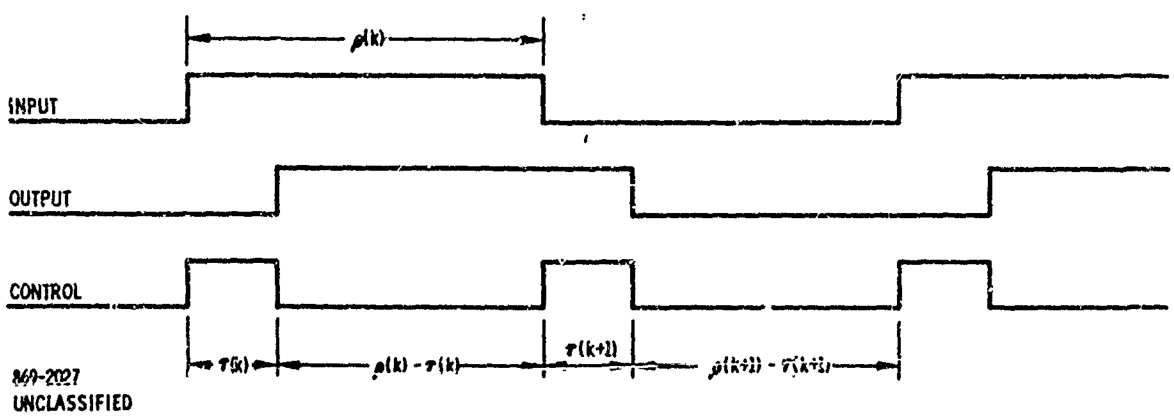

Figure 5-24. Waveforms for Digital Phase Lock Loop

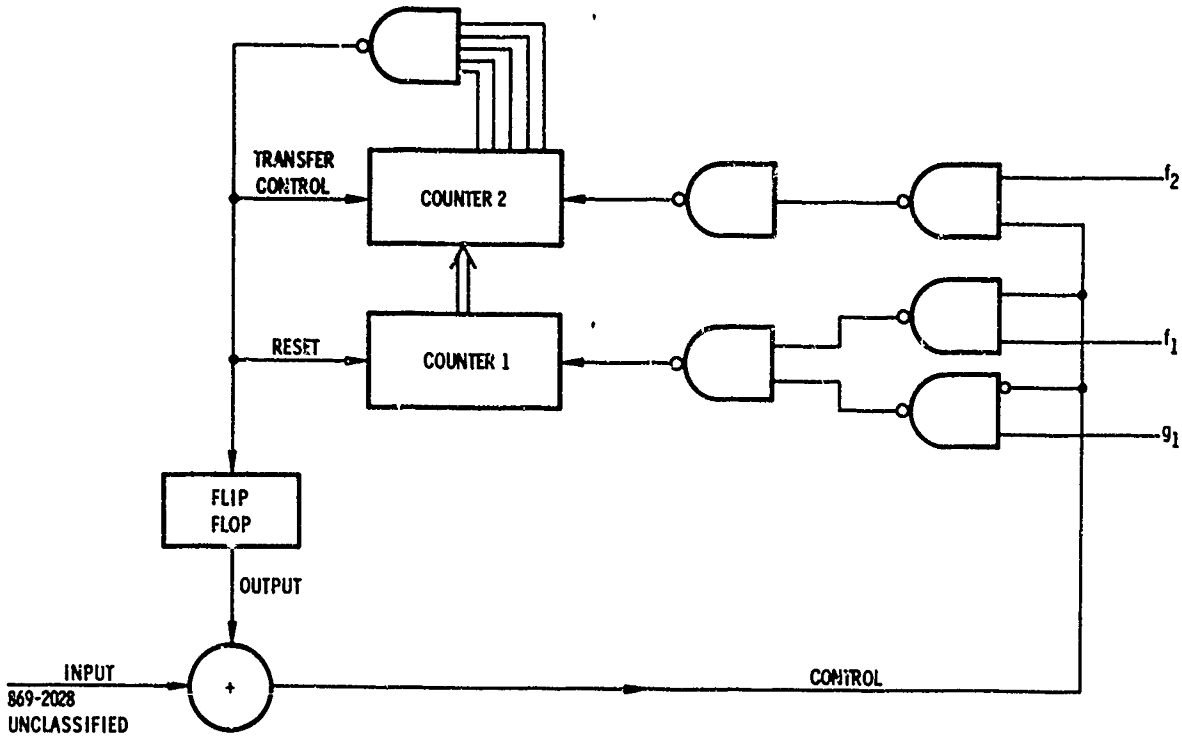

Figure 5-25. Second Order Digital Phase Lock Loop

$$f_1 = \frac{4M f_m \cos(\alpha)}{\cos(\alpha) - \sin h(\alpha)}$$

$$f_2 = \frac{2M f_m \exp(\alpha)}{\cos(\alpha) - \sin h(\alpha)}$$

$$g_1 = \frac{2M f_m \exp(-\alpha)}{\cos(\alpha) - \sin h(\alpha)}$$

$$\alpha = \frac{\pi f_n}{\sqrt{2}\, f_m}$$

where f_m is the average value of the input frequency and M is the decoded count of counter two that causes an output. The value of M is determined by the maximum amount of output phase jitter that is tolerable. The peak phase jitter can be calculated from

$$\frac{\tau(k)\ \text{peak}}{\tau(k)} = 1 + \frac{4}{M - g_1/f_m}$$

The second order PLL can be extended to a Costas loop as shown in Figure 5-26.

To demonstrate the ease at which a digital phase lock loop of the above type can be implemented, Figure 5-27 shows a block diagram of a second order PLL, where each block represents an integrated circuit that is available off-the-shelf today. The PLL requires 9 flat packages mainly because the specific functions required have never been put on a single chip. This implementation assumes the existance of f_1, f_2, and g_1. Because of the availability of the DM 7563 counter it was used throughout the design, but if a custom integrated circuit were built then a more simple counter design would be used. In addition two counter packages are shown in cascade which give a maximum count of $2^8 = 256$; a count this high would probably not be required.

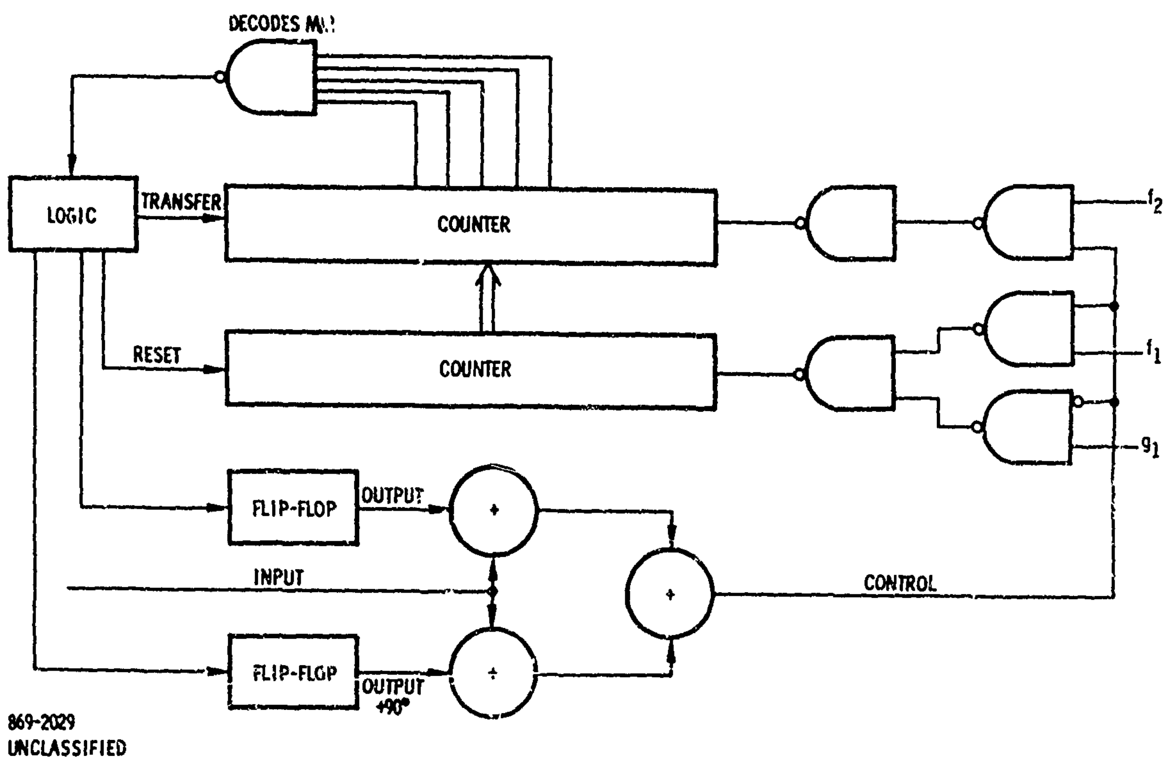

Figure 5-26. Digital Costas Loop

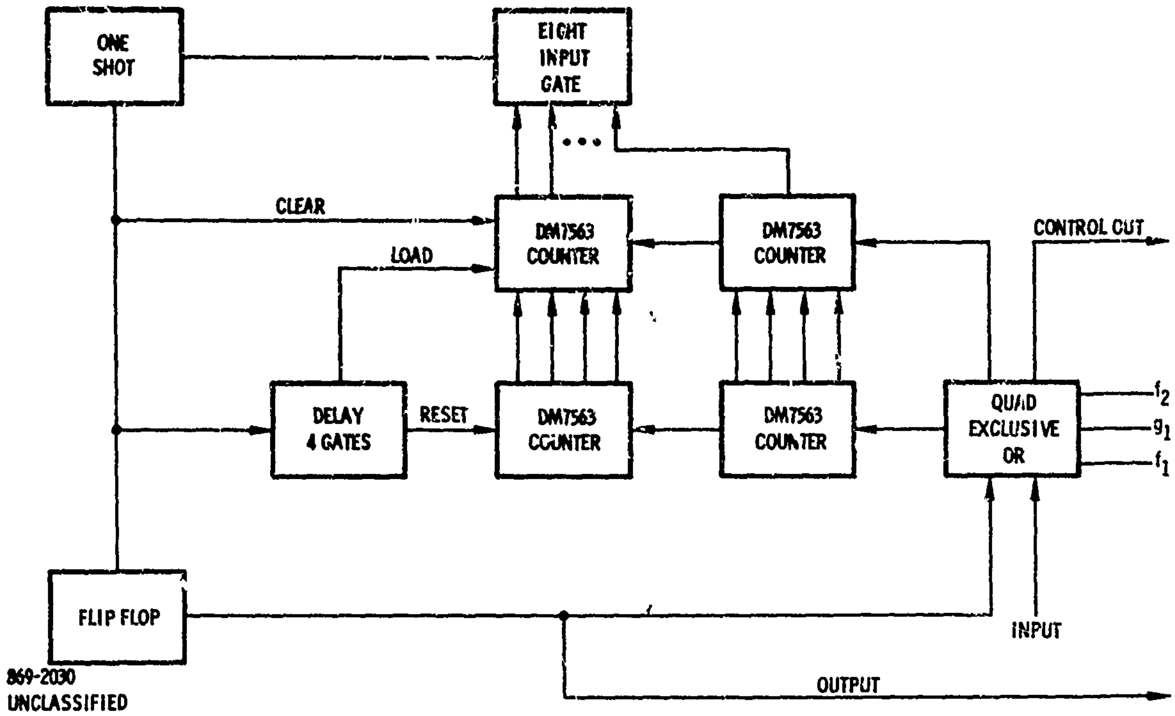

Figure 5-27. Digital Phase Lock Loop Implemented with Off-the-Shelf Integrated Circuits

If a custom integrated circuit were desired and the maximum count of 64 were sufficient, this digital PLL could be implemented on a single chip today. The complexity of this device is equivalent to about 14 flip-flops and 29 gates. This complexity is considerably less than the Model 3101-64 bit Bipolar Scratch Pad Memory made by Intel Corporation. There are very complex MOS circuits but due to the need to count at relatively high rates (i.e., about 64 times that of the input rate) the MOS complex arrays must be ruled out. If one were to produce a digital PPL today with minimum cost, minimum risk, and reasonably small size, the job should be done with three integrated circuits. The Motorola XC 177 takes care of the gates and two custom flip-flop chips would be required, but there is no doubt of obtaining them since 8 flip-flop circuits, on a single chip, are available from several manufacturers.

5.2.4.2.3 Sequential Decoder Implementation Study

This section describes the results of a preliminary study of the implementation of a sequential decoder potentially capable of operating at high data rates* (in excess

*This work was initiated originally in connection with Contract #DAAB07-68-C-0263 for U.S. Army Satellite Communication Agency.

of 1 megabit/second) as well as at 2400 bps for CNI application. The decoder uses a rate 1/2, systematic code, and accepts the raw binary digits from the demodulator. The algorithm is given in Appendix I of Volume II, and the performance realized in computer simulations is described in Section 7.3.2 of Volume II.

A decoder of the size under consideration requires the use of semiconductor memory, since the decoder is not large enough to realize the economies of scale for core memories. One feature of the above algorithm is that it allows specialization of the memory. A gross block diagram of the decoder is shown in Figure 5-28. The received information bits are shifted into an encoder that is a duplicate of the encoder used in the transmitter. The check bits generated by this encoder are mod-2 added with the received check bits. The result, called the syndrome, contains all of the information required by the decoder to detect errors in the received information bits. The syndrome is then entered into the random access memory (RAM). The information bits, which are not needed in the decoding process are delayed by an amount equal to the delay through the random access memory. The output of the RAM, which contains 1's in the error locations, is mod-2 added with the delayed information bits, thus accomplishing the correction of errors.

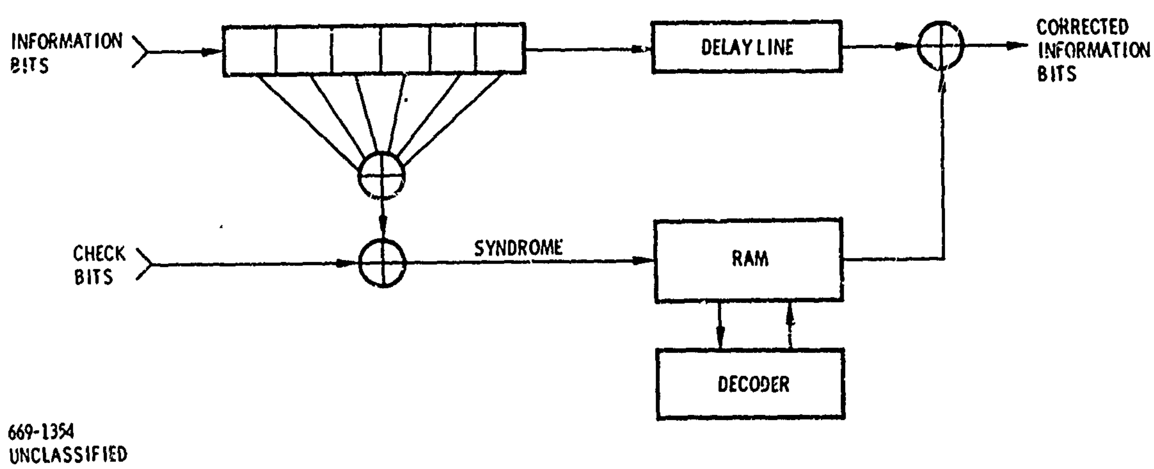

Figure 5-28. Sequential Decoder Block Diagram

The delay line may be implemented with MOS shift registers which offer considerably lower cost and higher density than fast random-access semiconductor memories. For example, dual 100-bit MOS shift registers are priced at \$30 or 15 cents/bit, while 16 bit RAM's in MECL II are priced at \$12 or 75 cents/bit. Thus, this decoder organization provides both a significant cost reduction with a significant package count reduction.

The algorithm is also designed to accomplish a maximum amount of computation each clock cycle in order to obtain a greater effective speed factor in the decoder. A tree diagram of the decoder possibilities is shown in Figure 5-29. If the present search node is 2, the decoder computes the best of the 6 other nodes and moves directly to that node. The Fano algorithm makes available only nodes 1, 4, and 5 as possible moves from node 2. This capability increases the effective speed factor of the decoder.

Another improvement is the mechanization of the encoders. As shown in Figure 5-28, the logic signals must propagate through $\log_2 (K/2)$ logic delays, where K is the constraint length, thus slowing the operation of the encoder. This is especially critical in the hypothesis encoder used in the decoder, since it adds directly to the time required to compute one node. A preferable implementation uses the impulse response of the encoder, which results in only one logic delay, thus speeding up the decoder computation rate.

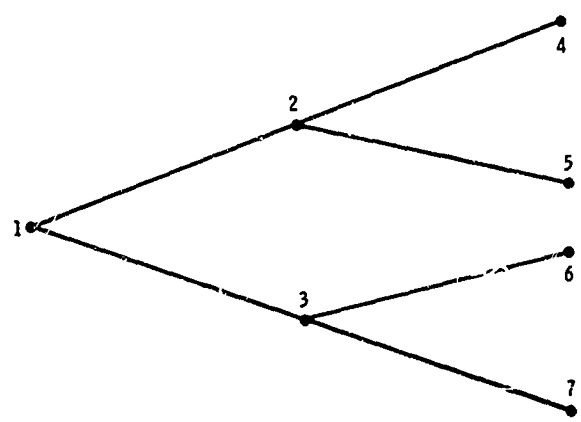

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Figure 5-29. Tree Diagram

A more detailed block diagram of the decoder is shown in Figure 5-30. The received information bits are passed through an encoder. The resulting check bits are mod-2 added, producing the syndrome. The syndrome is then buffered into the random-access memory. The error corrected syndrome is buffered out of the RAM and mod-2 added to the delayed "dirty" information bits producing "clean" bits for output. The RAM is organized into L-2 bit words. Four 2-bit words are read in and out in parallel to provide an effective speed up of four in memory access time. The memory address is selected from one of two counters. One counter contains the address of the input-output, while the other contains the address of the decoder.

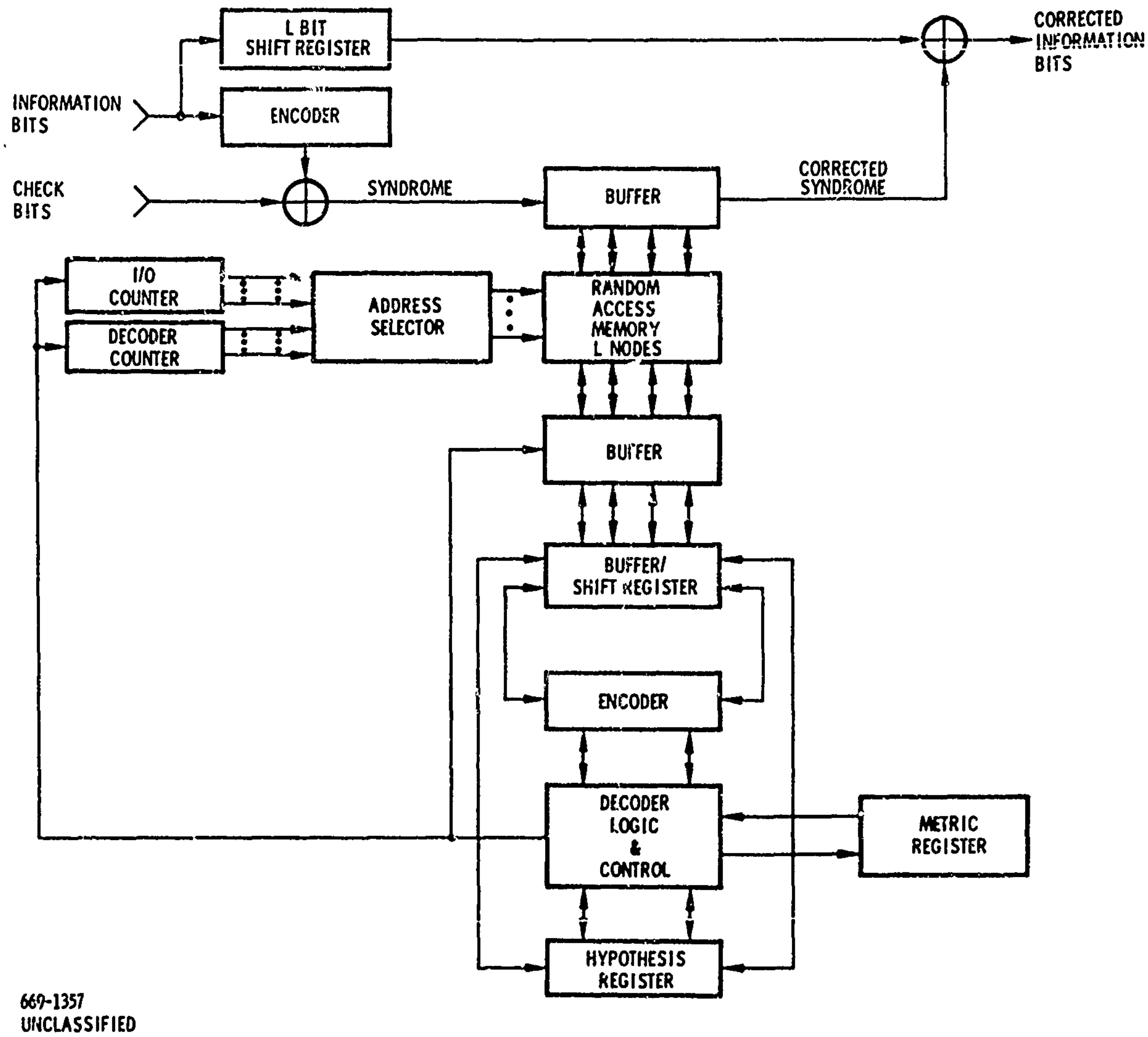

Figure 5-30. Sequential Decoder Block Diagram

The I-O counter is incremented by one every fourth information bit, and a I-O read-write cycle is stolen from the decoder. The decoder counter is an up-down counter that is incremented by one each time the decoder moves ahead or behind four nodes.

There is a two-level buffer on the decoder side of the RAM. This is to enable the decoder to proceed at maximum speed in either direction or reverse itself at any time without having to wait for an extra memory cycle. The second level buffer is also a right-left shift register. The syndrome in this buffer is shifted into an encoder. The decoder logic determined the presence of errors and inserts a "one" in the hypothesis register when errors are located. The decoder logic also computes the new value of the metric, the tilted distance function, and stores this in the metric register. This logic also decides whether to search forward or backward and raises and lowers the threshold when needed.

It should be noted that there is no separate back-up buffer; it has been included in the RAM. This was done because the RAM, which is constructed of MECL II circuits, is sufficiently fast that a separate back-up buffer is unnecessary, thus resulting in fewer packages.

It is estimated that the decoder logic and control together with the metric register can be implemented in about 40 flat packs of Motorola MECL II and MECL III. This figure is independent of buffer size and encoder constraint length. For a constraint length of 32, the implementation requires a total of about 250 flat packs for the encoder and the RAM for 512 nodes of storage, including associated buffers and address counters. It is estimated that the above decoder can be operated at a clock rate for computing as high as 50 MHz; hence, is more than capable of CNI data rates.

APPENDIX I

PAIRED ECHO INTERPRETATION OF DISTORTION

The guidelines for 100 MHz and $\pm$ 3.5 degrees were obtained by considering two cases: amplitude distortion only and phase distortion only. The approach is briefly described below and is used as a guideline only. In actuality the process is complex and involves both phase and amplitude distortion.

I-1. AMPLITUDE DISTORTION

To a first order approximation distortion due to amplitude can be viewed independently as follows. Given a characteristic as shown in Figure I-1 where the phase is proportional to frequency and the amplitude characteristic is sinusoidal

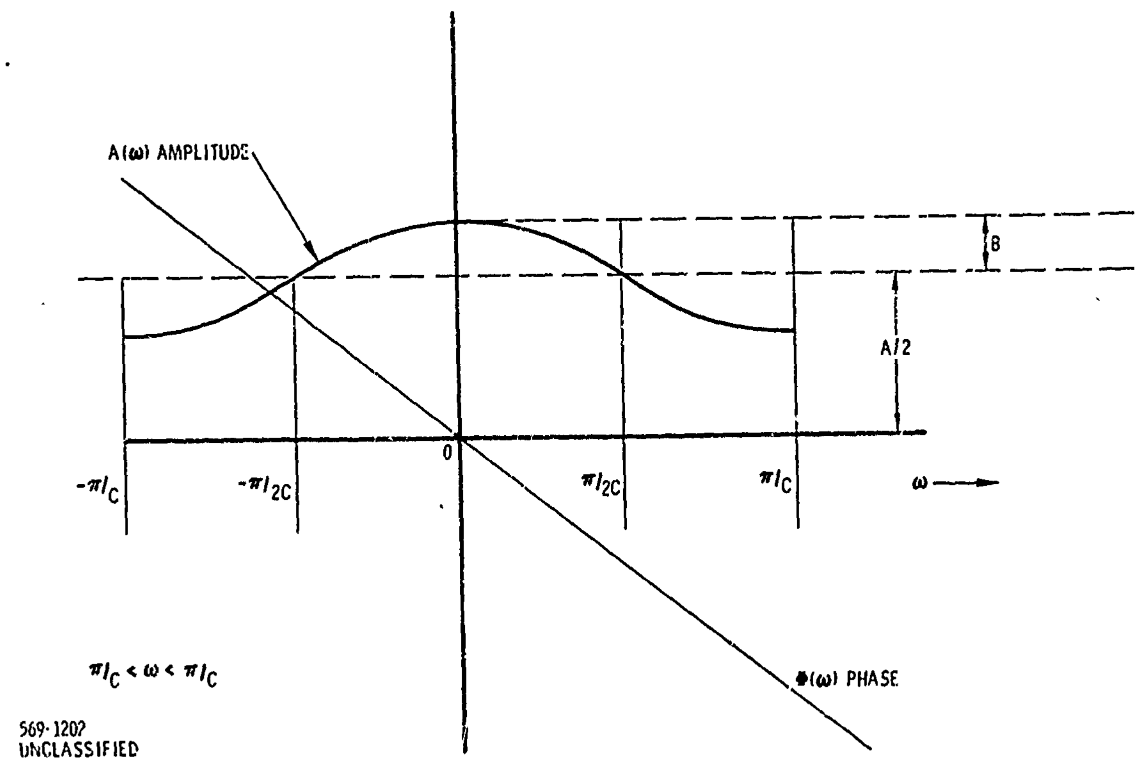

Figure I-1. A Nonuniform Amplitude Transmission Characteristic

$$\Phi(\omega) = -b_o\omega \tag{1}$$

$$A(\omega) = \frac{A}{2} + \beta \cos \omega c \tag{2}$$

It can be shown that the output*

$$\overline{G}(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(\omega) \left(\frac{A}{2} + \beta \cos \omega_c\right) e^{+j(\omega t - b_o\omega + \phi(\omega))} d\omega \tag{3}$$

is the desired signal $G(t-b_o)$ plus two signal echoes; i.e.,

$$G(t) = \frac{A}{2}\ G(t-b_o) + \frac{B}{2}\ G(t-b_o+c) + \frac{B}{2}\ G(t-b_o-c) \tag{4}$$

The ratio of the echo to the signal B/A, therefore, by confining the ratio, the distortion becomes negligible. For example, a ratio of B/A = 0.03 for the case of A/2 = 1 results in a value of B = 0.06. Thus the amplitude variation

$$\frac{A+B}{A} = 1.06 = 0.5\ \text{db}$$

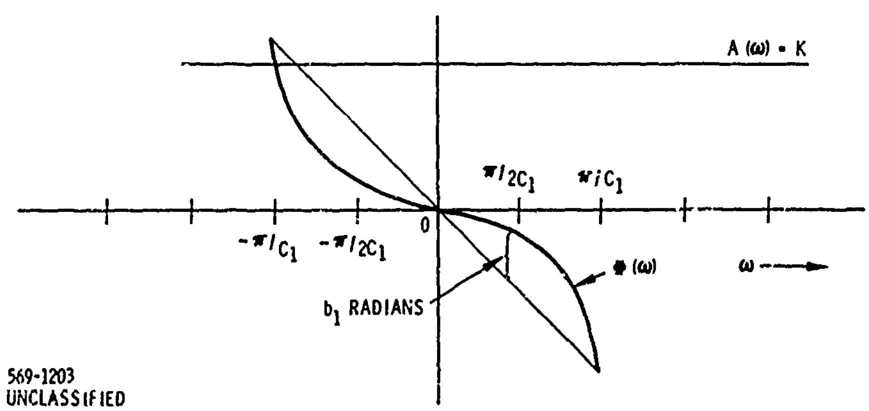

Figure I-2. A Nonlinear Phase Transmission Characteristic

*S. Goldman, "Frequency Analysis, Modulation and Noise," pp. 103.

I-2. PHASE DISTORTION

For the case of constant amplitude and phase variation from linear (b_1 in radian) as shown in Figure I-2, the effect of phase variation to a first order approximation can be seen.

$$A(\omega) = K \tag{5}$$

$$B(\omega) = -b_0\omega + b_1 \sin c_1\omega \tag{6}$$

Here again the output $\overline{G}(t)$ can be shown as a signal, undistorted, plus echoes where

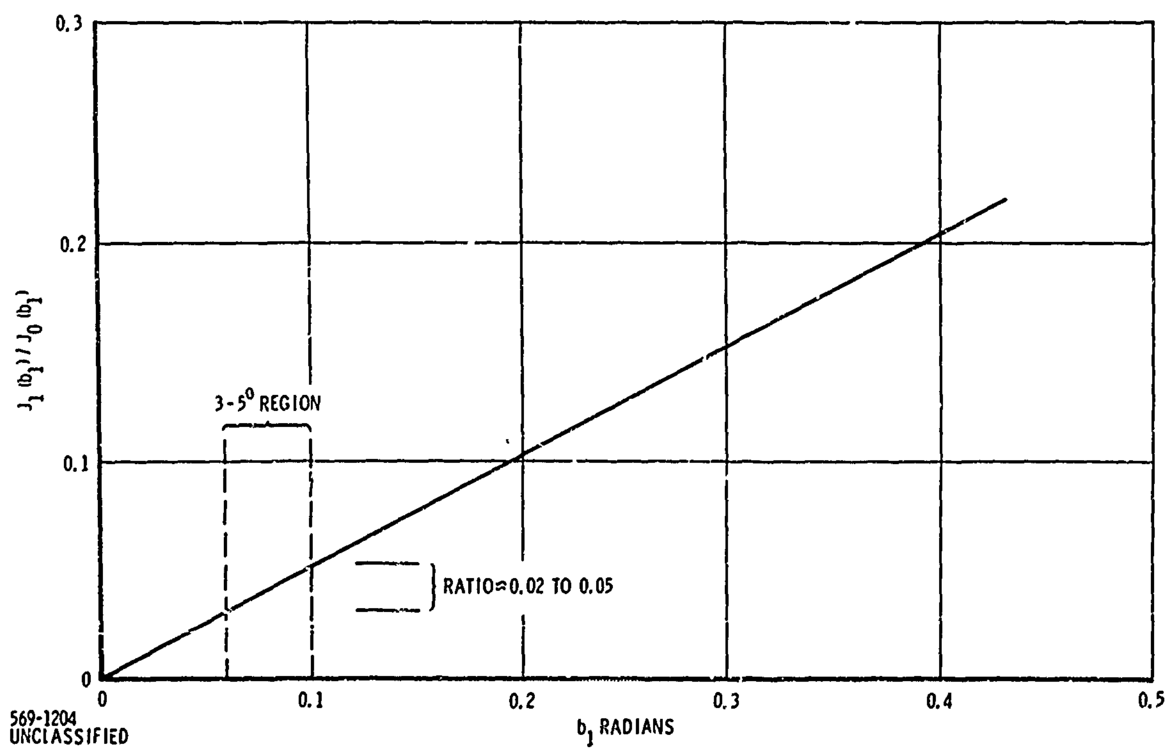

Figure I-3. Ratio of $J_1(b_1)/J_0(b_1)$ Versus b_1

$$\tilde{U}(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(\omega)\, K\, e^{j(\omega + \phi(\omega) - b_o\omega + b_1 \sin c_1\omega)}\, d\omega =$$

$$K\left[J_o(b_1)\, g(t-b_o) + J_1(b_1)\, G(t-b_o+C_1) - J_1(b_1)\, G(t-b_o-c)\right] \tag{7}$$

the echoes here are of opposite polarities and given the proper relationship between c_1 and the pulse width (of a pulse train, for example) could conceivably cancel. To minimize the phase distortion for the above case, the ratio of echo to signal $J_1(b_1)/J_o(b_1)$ must be minimized. Figure I-3 shows the ratio as a function of the excursion in radians. For $0.02 < b < 0.05$, the variation from linearity must be 3-5°.

APPENDIX II
CIRCULATOR CONSIDERATIONS

The best technique for separating the input and output signals is through the use of a circulator. The simplest form of the circulator coupled with a tunnel diode amplifier is shown in Figure II-1. Typical circulator parameters for a single junction are: (1) VSWR 1.2, (2) forward loss 0.2 db, (3) reverse loss 20 db. In order to examine the behavior of the various types of circulator coupled tunnel diode amplifiers (Figures II-1 - II-4), let us assume that the above circulator parameters hold for each junction and that the tunnel diode amplifier has 17 db gain. By applying these parameters to evaluate the effects of various signal paths for the four amplifiers we obtain the results presented in Table II-1. It is obvious that the 3-port circulator coupled tunnel diode amplifier will be satisfactory only at very low gain or where the source and load matches are very good. The 4-port circulator with the tunnel diode amplifier on the first junction has good load stability and with the tunnel diode amplifier on the second junction has good source stability. The 5-port circulator on the second junction has good source stability. The 5-port circulator tunnel diode amplifier has good load and source stability. It is best for most applications; however, it is slightly larger and does have 0.2 db higher noise figure than the first two because of the extra junction in front of the tunnel diode circuit. Another factor to consider is out-of-band oscillation. Any condition within the amplifier that leads to a loop gain through the tunnel diode amplifier of unit at any frequency will cause oscillation.

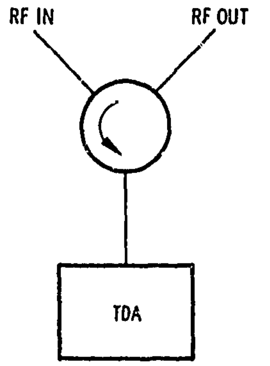

Figure II-1. 3-Port Circulator TDA

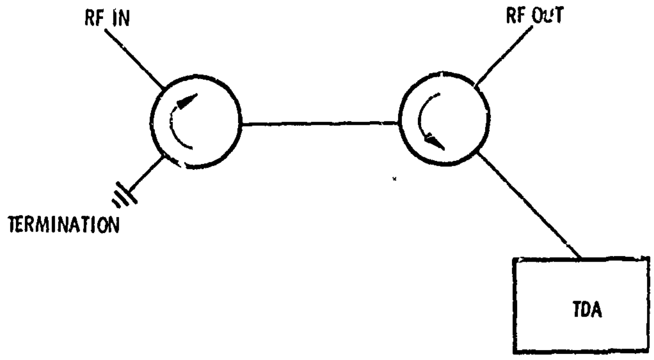

Figure II-3. 4-Port Circulator TDA On Second Junction

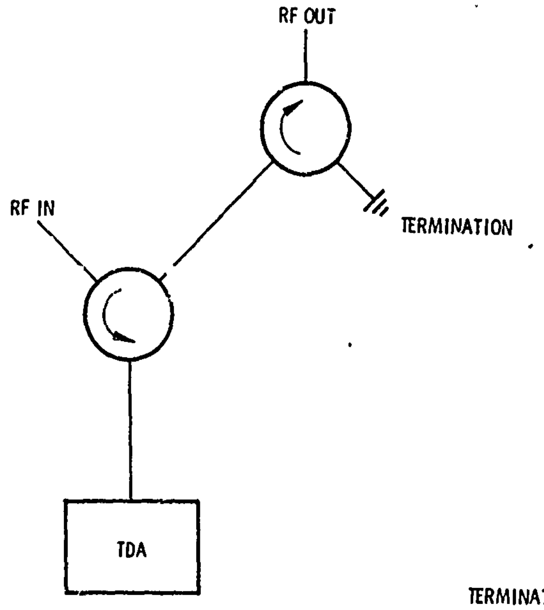

Figure II-2. 4-Port Circulator TDA On First Junction

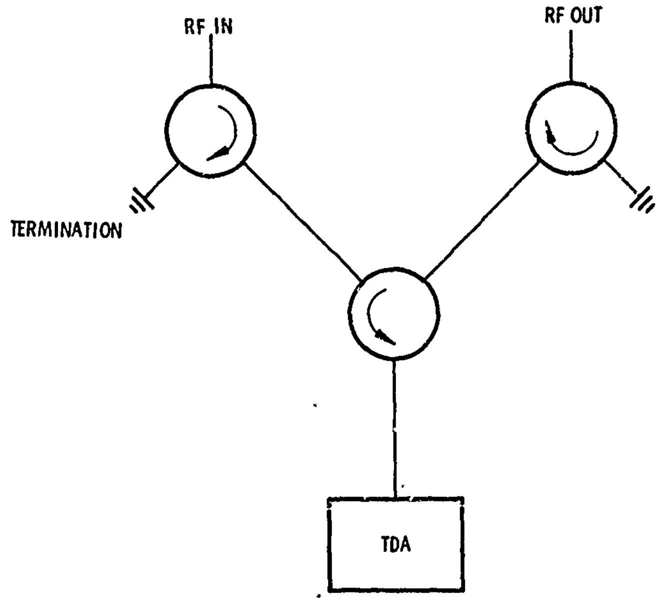

Figure II-4. 5-Port Circulator TDA

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Table II-1. Circulator Considerations

17 db Gain Tunnel Diode Amplifier (Nominal)	Input VSWR with output Terminated	Output VSWR with input Terminated	Δ Gain (max) Source and Load VSWR 3:1	Δ Gain (max) Source and Load VSWR 1.5:1	Δ Gain (max) Source VSWR 3:1 Output Terminated	Δ Gain (max) Load VSWR 3:1 Input Terminated
3-port Circulator Tunnel Diode Amp	6:1	6:1	Oscillates	5.0	6.0	6.0
4-port Circulator with Tunnel Diode Amplifier on First Junction	6:1	1.15:1	.0	2.5	3.0	.7
4-port Circulator with Tunnel Diode Amplifier on Second Junction	1.15:1	6:1	6.0	2.5	.7	3.0
5-port Circulator Tunnel Diode Amplifier with Tunnel Diode Amplifier on Second Junction	1.15:1	1.15:1	3.0	.35	.7	.7

SECTION VI
DEMONSTRATION CONCEPT

The majority of the effort expended during the study was devoted to topics within Task A001, which defined a preferred full-capacity waveform for ultimate CNI application. The studies and rationale leading to selection of the FH/PN/TH full-capacity waveform have been presented in the preceding sections. This section develops a recommended demonstration concept suitable for immediate implementation within the state of the art to indicate feasibility of the full-capacity preferred waveform and validity of the general integration concept of CNI.

6.1 SIGNAL DESIGN PARAMETER SCALING (TASK A002)

The purpose of Task A002 is to derive a scaled down version of the full-capacity CNI signal design which can be readily and economically implemented in order to provide a test bed for the experimental evaluation of the CNI concept. It is important to emphasize that the efforts in support of this task were primarily motivated towards the functional development of a CNI demonstration transceiver with relatively little regard for the potential operational problems inherent in a fully deployed system. However, in order to sensibly scale the parameters of the full-capacity signal design, it was necessary to hypothesize a limited set of nominal system constraints.

In regards to postulating an operational environment which provides the overall rationale for sizing the parameters, it was decided to restrict consideration to an organized hypothetical satellite mode. This situation allows an analysis of a state of the art down link power budget without having to formulate a system operational doctrine or a detailed system organization. The only essential feature which this approach does not take into account is the near-far problem associated with a direct (nonsatellite) mode of operation. However, provision can be made to demonstrate experimentally the performance of the CNI waveform in a simulated near-far environment if it is deemed necessary.

Following the general sizing of the signal parameters, consideration is given to a functional design of a transceiver which will demonstrate the attributes of the preferred signal design. At this point in time it would appear sensible to restrict

the demonstration to the controlled environment, and in addition (at least initially) restrict the equipment required to a pair of modems. This obvious simplification which will significantly reduce hardware costs associated with the demonstration can be justified on a number of counts. For example, it avoids having to make what appears to be at this time a very arbitrary choice with regard to operating frequency. Secondly, the RF equipment should not seriously degrade the signal processor's performance. In a similar vein, rather than processing four individual navigation signals as would be the case for a full-capacity system, the demonstration will be confined to the time of arrival measurement of a single navigation signal. Since the basic requirement of the demonstration is to perform multiple signal processing, both a voice and a data link are included in addition to the navigation signal. However, the Statement of Work calls out an RF implementation of the demonstration transceiver which could enable testing between ground and an airborne platform. Hence, functional design will enclude the RF at roughly 1.5 GHz, so as to meet the stated requirements.

The preferred full-capacity signal structure is described generically as a FH/PN/TH hybrid. We intend to impose a phase coherency requirement on the frequency hopping in order to demonstrate feasibility of very accurate time of arrival measurements (i.e., NAV function) consistent with total bandwidth occupancy. It is convenient from an implementation point of view to select the total number of frequency slots available to the system to be equal to some power of two. Here we will arbitrarily choose 32 frequency slots each of which are equally likely and one of which can be occupied by a modulated PN carrier at a given instant in time. It is also very convenient to select a basic time slot size which is equal to or some submultiple of the frequency hop dwell time. In general then a particular transmission burst may occupy several of these elementary time slots in order to conduct reliable processing at the receiving terminal.

For our purposes here it is sufficient to characterize the hypothetical (near-term) satellite system in terms of a down link power budget as shown in Table 6-1. This clearly represents the limiting case where the receiving terminal consists of a typical high performance aircraft without a directive antenna. The path loss calculation was based on the arbitrary choice of an L-band operating frequency and a maximum slant range of 25 x 10^6 meters. We note that for this hypothetical situation the received signal power to noise power density (per Hz) ratio is 50 db.

Table 6-1. Down Link Power Budget Summary

Transmitter Power (30 watts)	15 dbw
Transmitter Antenna Gain	20 db
Path Loss $(\lambda/4\pi D)^2$	-190 db
Misc. Losses	- 3 db
Receiver Antenna Gain	0
Received Power	-158
Noise Power Density	-208

Because of the inherent TDMA nature of the signal design, it is necessary to determine a minimum time slot size. It can be readily shown that the minimum slot size is given by the expression

$$T_s = \frac{E/N_o}{P/N_o} \cdot M$$

where

E/N_o = the required energy per bit to noise density ratio

P/N_o = the received signal power to noise density ratio

M = the number of bits in a time slot

It is evident that the slot size depends on the modulation scheme through the E/N_o ratio. For the demonstration system, we propose to employ DPSK as the modulation form. To satisfy the voice communication requirements, we suggest the use of a 2400 bps vocoder which we assume can be made available for the demonstration. If we take as an appropriate threshold a 1% bit error probability, an E/N_o of approximately 6 db is required at the receiver. The corresponding minimum time slot size is then of the order of 40 microseconds per bit. If we now arbitrarily select a frame time of approximately 1 second, we can transmit a burst of 2400 bits of vocoded voice in 160 milliseconds at a burst rate of 15.4 kbps. An additional portion of the 1 second frame must be reserved for digital data. For this reason, we consider as appropriate a simulation of a data source which provides data at a 4.8 kbps rate.

Without coding a bit error rate of 10^{-3} is adequate for the purpose at hand. This requires an E/N_o of approximately 8 db thus establishing a minimum slot size of 62.15 microseconds per bit. It is now possible to transmit a frame of digital data in 320 milliseconds at a burst rate of 15.4 kbps. The remainder of the frame (i.e., 544 milliseconds) is devoted to processing the navigation signal. It is convenient to select for the demonstration system an elementary slot size of 32 milliseconds so that five slots are required for a complete burst of vocoded voice and ten for the digital data making a total of 15 equal length time slots. The remainder of the frame consists of essentially 17 time slots all of which are devoted to the navigation signal. For the demonstration system, we propose to reserve the last 17 time slots of each frame for the navigation signal. However, we will pseudorandomly permute the first 15 time slots of each frame. A particularly simple way to implement the permutation of these time slots is to inject four bits from a long PN sequence into a 4-stage linear feed back shift register at the beginning of each frame. The initial state of the register at the start of a frame period will then be a random number between 1 and 15 (the zero vector is suppressed). Shifting the register once every 32 milliseconds for a period of 480 milliseconds generates a sequence of orthogonal numbers which can be used to identify for transmission the fifteen 32 millisecond segments of vocoded voice and digital data in storage. Of course, much more sophisticated permutation techniques are available but they are considerably more costly to implement.

From an implementation point of view it is convenient to select a frequency hopping rate of approximately 31 hops per second. More precisely we suggest a frequency hop dwell period of 32 milliseconds which corresponds to an elementary time slot. The implementation consists of selecting a 5 bit sequence from a long PN code every 32 milliseconds so that each of the 32 frequency slots appear with equal probability. A 9.6 MHz bandwidth is subdivided into 32 subbands each one being 300 KHz wide.

6.2 FUNCTIONAL DESIGN OF TRANSCEIVER/PROCESSOR (TASK A003)

A functional block diagram of the transmitter section is shown in Figure 6-1. At the beginning of each frame, the buffer storage, which consists of 15 registers, contains a total of 7373 bits. Five of these registers correspond to 5 time slots of 2.4 kbps data while the remaining 10 registers correspond to 10 time slots of 4.8 kbps

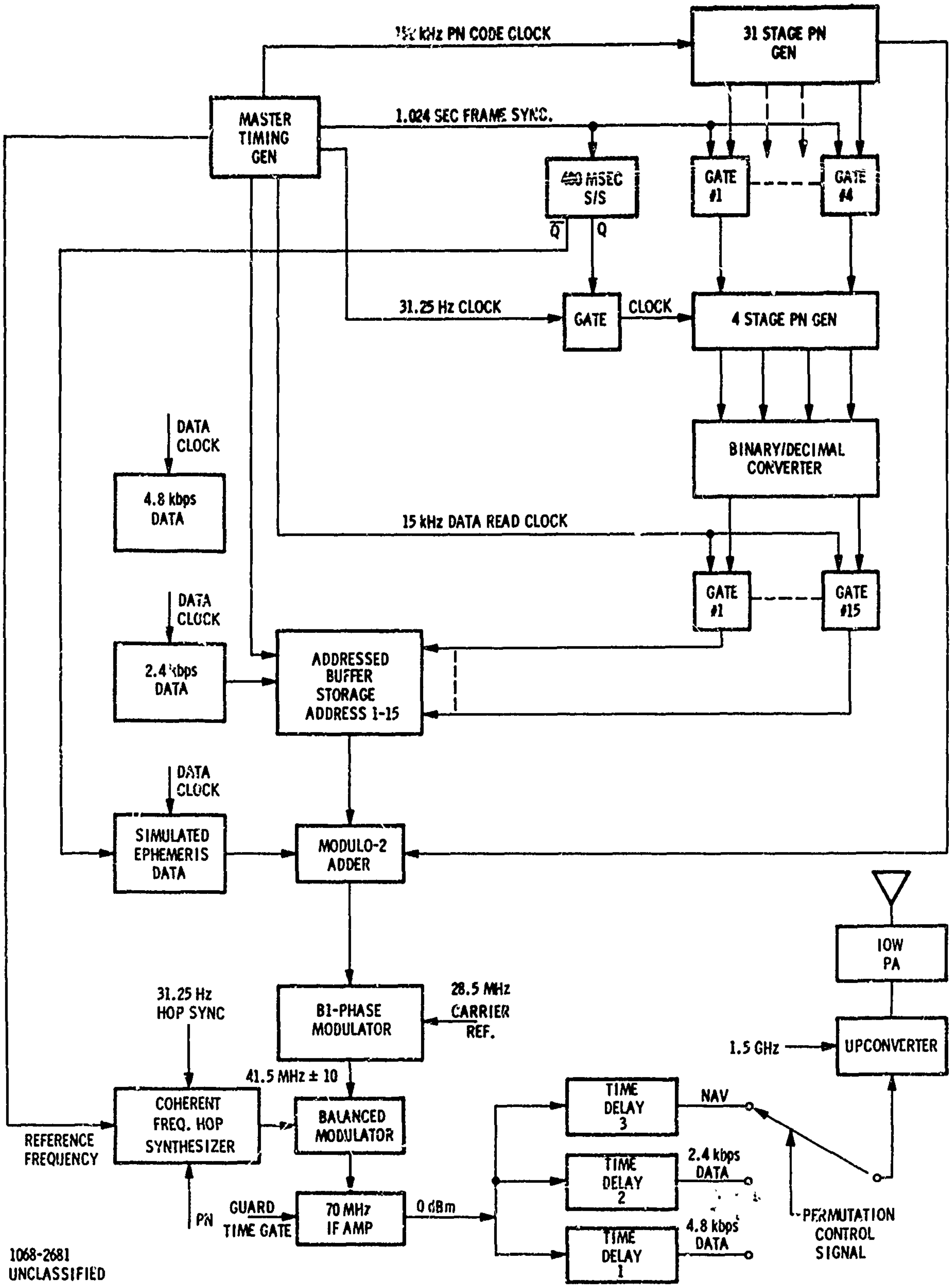

Figure 6-1. CNI Demonstration Transmitter Section

data. The four bit sequence selected from the 31 stage PN generator initiates an orthogonal pattern of fifteen numbers which are used to match registers to the sequence of time slots. In each slot, the data contained in the selected register is modulo-2 added to the PN code and the resulting digital signal biphase modulates a fixed carrier at 28.5 MHz. During each time slot, the coherent frequency hopping synthesizer randomly selects one of 32 possible frequencies in the range from 36.5 MHz to 46.5 MHz, to translate the PN modulated carrier to a center frequency of 70 MHz. If the time slot contains a burst of 4.8 kbps data, the output switch automatically switches to Time Delay #1 which simulates a propagation delay. On the other hand, if a burst of 2.4 kbps or a NAV transmission occurs, a different simulated propagation delay is inserted in the signal transmission path. In this manner we simulate three separate signal emitters with a single transmitter section. Of course, the switching function is controlled by the orthogonal number generator so that as we permute the time slots it is guaranteed that each transmission burst passes through the appropriate simulated propagation delay. If an RF interface is required an additional module containing a 1.5 GHz synthesizer, an up converter and a 10 watt PA will be provided.

A simplified functional diagram of the signal processor is shown in Figure 6-2. Observe that the signal processor can be functionally divided into three basic subsystems. First, a digital matched filter (DMF) noncoherent detection scheme is employed to acquire initial synchronization at the beginning of each time slot. To facilitate proper operation of the DMF, data modulation does not appear on the received signal at the beginning of each time slot for a time interval corresponding to the maximum time uncertainty. During the process of initial acquisition, the second subsystem (i.e., the Costas loop) which provides the means for synchronous data demodulation and estimating range rate is disabled. Similarly, the third subsystem (i.e., the coherent frequency hopping tracking loop) which provides the precision tracking function for estimating range is disabled by virtue of the fact that the Costas loop filter is open. Summarizing to this point, we note that the three basic subsystems which characterize the proposed implementation of the signal processor consists of (1) noncoherent sync acquisition DMF detector, (2) Costas loop coherent data demodulator, and (3) coherent FH precision range tracking loop.

We might point out that although PN tracking has not been included in the proposed demonstration implementation at this time, it may be included in a full capacity system as a valuable aid to the precision range tracking loop during acquisition or

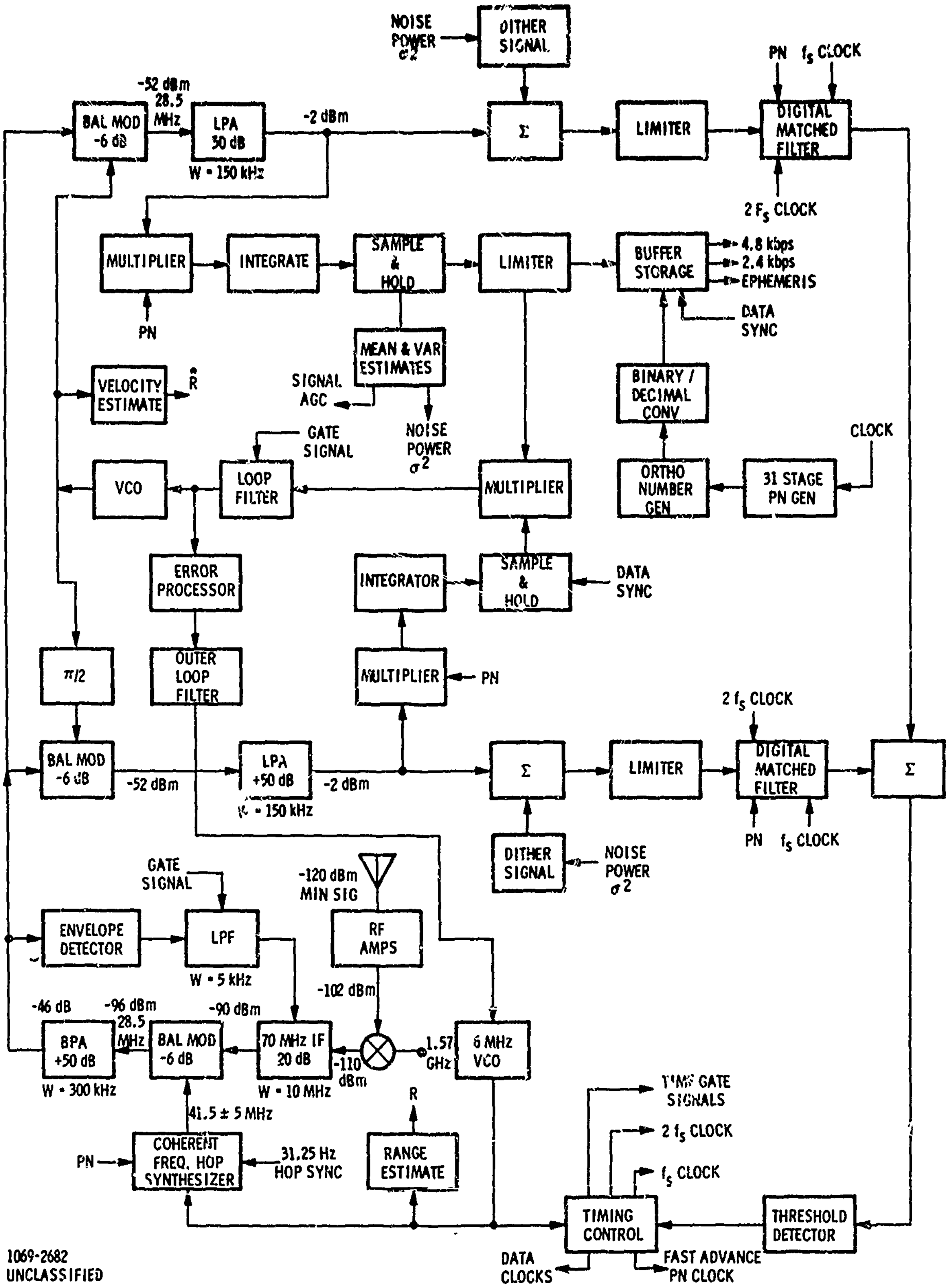

Figure 6-2. CNI Demonstration Signal Processor

even as suitable by itself. Another comment relative to the difference between the demonstration and full capacity systems is concerned with coding. For instance, the use of error correcting codes as proposed for the full capacity system would relieve the problem of having to synchronize very reliably within each time slot, provided the error control code with interleaving spans more than a single slot. This, of course, allows for a lower threshold value of E/N_0 than that required for the less complex demonstration system proposed here. Another point which needs to be mentioned is concerned with the implementation of the matched filter. In particular, we have not implemented the DMF to handle the doppler situation. The reason for this is based simply on the fact that if the demonstration signal parameters are scaled upwards to represent the full capacity signal parameters the DMF integration time is reduced to the point where Doppler does not appear as a problem.

Referring to Figure 6-2, the signal received by the signal processor is first passed through a 10 MHz wide IF amplifier centered at 70 MHz. The output, which is still a frequency hopped signal, is multiplied by a local oscillator which is frequency hopped according to the same pattern used at the transmitter. The multiplier or balanced modulator output is then passed through a bandpass amplifier whose bandwidth is at least 300 KHz. The output of this filter is essentially the PN modulated carrier plus noise which is applied to both an active envelope detector for fast AGC during signal acquisition and to quadrature detectors. Note that once the DMF detector indicates sync has been achieved this method of AGC is changed to one which operates on a measurement of signal strength as will be described subsequently. The outputs of the quadrature balanced modulators are low pass filtered (minimum bandwidth of 150 KHz) which in the absence of noise provides signals [DATA x PN] cos ϕ in the inphase channel and [DATA x PN] sin ϕ in the quadrature channel. Of course, as we have mentioned previously there is no data modulation present during the initial sync acquisition interval. As described in Section 7.5.2 of Volume II, a dither signal is added to the low pass filter output in order to minimize the signal to noise ratio degradation due to the nonlinear operation of the limiter. In order to synthesize the appropriate dither signal, it is necessary to measure the noise power σ^2. This can be done by computing the variance of the output of the integrate and dump filter in the data channel. We might note at this time that once sync has been acquired the mean value of the I&D filter output is proportional to the signal energy and can be used as a control signal for AGC purposes during the remaining portion of the time slot.

Returning to the DMF detector the output of the limiter is a sequence of plus and minus ones which in the absence of noise is the PN sequence. The limiter output is clocked into the signal register of the DMF at a rate of at least twice the code clock. The reference register contains a subsequence of PN bits which are used to weight the sequence in the signal register. The sum of the individually weighted stages in the signal register will be a maximum when the two registers contain identical sequences. The outputs of both the inphase and quadrature DMF's are combined and applied to a threshold detector which indicates with a pulse the alignment of the received code with the static subsequence in the reference register. Once this event occurs, the reference PN generator is activated in order to keep the signal and reference registers matched.

The length of the reference register effectively establishes the detection SNR. Recall that an E/N_0 of 8 db is required for communication purposes. For a one bit integration time and taking into account the limiter loss yields an E/N_0 of approximately 3 db which is far too small for reliable sync detection. If we choose a detection SNR of approximately 12 db which corresponds, on a per trial basis, to a detection probability of 0.96 and a false alarm probability of 0.33×10^{-3} the integration time must be increased to about 8 bit times (i.e., 8/15 milliseconds). This establishes the minimum length of 80 bits for the reference register in the DMF.

Once sync has been acquired, the timing control generates a gate signal which closes the Costas loop filter allowing both the Costas loop and the precision range tracking loop to acquire and to begin tracking. The data demodulator first removes the PN code by correlation and the resulting data signal is matched filter detected and applied to the receiver's buffer storage. Here the data is temporarily stored in its proper location by means of the orthogonal number generator which controls the time slot permutations. The range measurement is made by first processing the Costas loop filter output to obtain an error signal which is proportional to the difference in adjacent frequency hopping tones, $\Delta\omega$, multiplied by the relative timing error. This error signal after multiplication by the sign of the frequency hop is filtered by a proportional plus integral control filter whose output controls the phase of the VCO which acts as the receiver's timing reference. A comparison of this VCO's output with that of a fixed time standard provides a measure of range.

The range rate measurement is made by the relatively standard technique of cycle counting the Costas loop VCO output, provided that coherent synthesizers are employed throughout, including the NAVSAT as well as the user receiver.

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13. ABSTRACT

This Final Report, presented in three volumes, describes a comparison of candidate spread spectrum waveforms and the selection of a preferred waveform to perform integrated communication, navigation, and identification (CNI) functions. Satellites are presumed available in appropriate orbits for global communication and navigation. A coordinated frequency hop/pseudonoise/time hop (FH/PN/TH) waveform is made considering such factors as efficient use of satellite ERP in the remote mode, multiple access of wide dynamic range signals in the direct mode, range and range rate measurement accuracy, initial synchronization, and equipment complexity for full capacity implementation in a nominal 100 MHz bandwidth. Since CNI system requirements are not presently known, the waveform choice has been made considering a postulated worst case environment based on future air traffic control requirements. Implementation of the preferred CNI waveform will depend on certain technology developments particularly in the areas of wide dynamic range receivers, phase coherent frequency hopping, high peak power pulse transmitters, and LSI digital devices. However, a demonstration concept can be advanced within the present state-of-the-art to illustrate the preferred waveform with scaled parameters. Volume I covers the concept formulation studies leading to the preferred waveform and demonstration concept while Volume II summarizes the detailed performance and operational analysis. Volume III presents navigation considerations for the enroute case.

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